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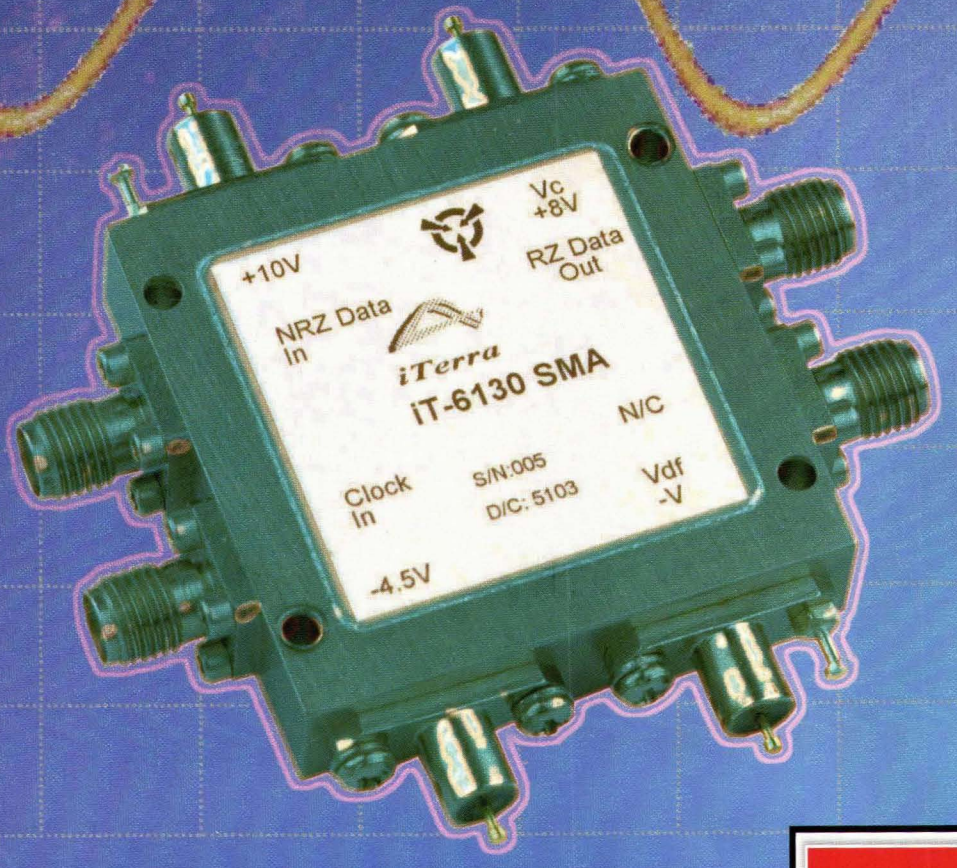
News
Sampling the latest ADCs/digitizers

Design Feature
Selecting ICs for remote keyless entry

Product Technology
CMOS fires GSM power amp

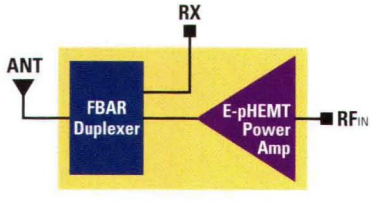
Module Melds 11.5-Gb/s NRZ/RZ Converter and Driver



Communications Issue

Two good

Agilent RF technologies make one great front-end solution



CDMA 1900 FEM Example Block Diagram

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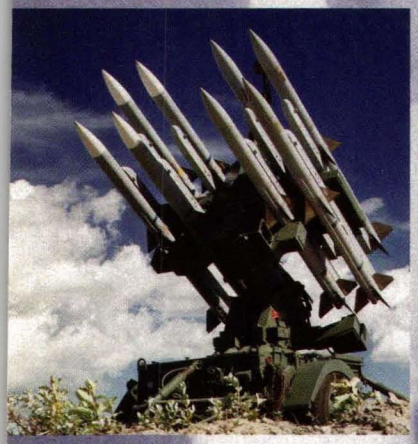

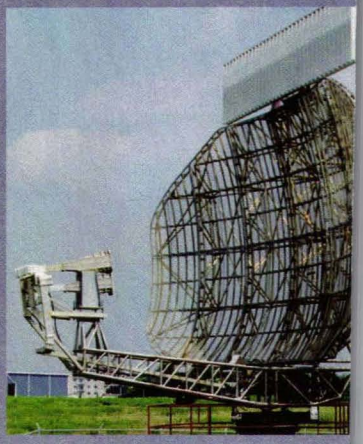

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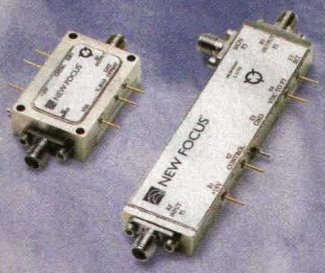

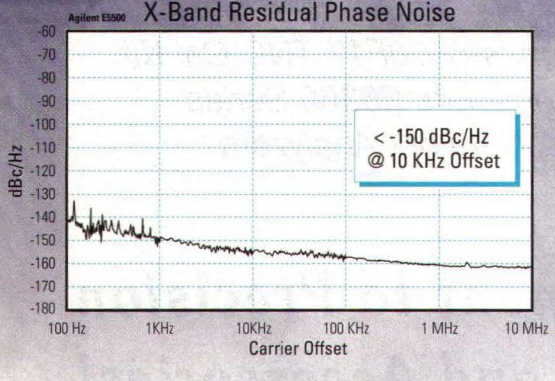

Broadband Power Amplifiers

Model	Freq. Range GHz	Gain dB min	N/F dB max	Flatness +/–dB	1 dB Comp. pt. dBm min	3rd Order ICP typ
JCA018-3000	2.0–18.0	25	6.0	2.0	23	28
JCA218-3001	2.0–18.0	25	6.0	2.0	25	30
JCA218-3002	2.0–18.0	25	6.0	2.0	27	32
JCA218-4000	2.0–18.0	30	6.0	2.0	23	28
JCA218-4001	2.0–18.0	30	6.0	2.0	25	30
JCA218-4002	2.0–18.0	30	6.0	2.0	27	32
JCA218-5000	2.0–18.0	35	6.0	2.0	23	28
JCA218-5001	2.0–18.0	35	6.0	2.0	25	30
JCA218-5002	2.0–18.0	35	6.0	2.0	27	32

Power Amplifiers

Model	Freq. Range GHz	Gain dB min	N/F dB max	Flatness +/–dB	1 dB Comp. pt. dBm min	3rd Order ICP typ
JCA12-P01	1.35–1.85	35	4.0	1.0	33	41
JCA34-P02	3.1–3.5	40	4.5	1.0	37	45
JCA56-P01	5.9–6.4	30	5.0	1.0	34	42
JCA812-P03	8.0–12.0	40	5.0	1.5	33	40
JCA1218-P02	12.0–18.0	22	4.0	2.0	25	35

Low Noise Amplifiers

Model	Freq. Range GHz	Gain dB min	N/F dB max	Flatness +/–dB	1 dB Comp. pt. dBm min	3rd Order ICP typ
JCA12-1000	1.2–1.6	25	0.8	0.5	10	20
JCA12-3001	1.0–2.0	40	0.8	1.0	10	20
JCA23-302	2.2–2.3	30	0.8	0.5	10	20
JCA34-301	3.7–4.2	30	1.0	0.5	10	20
JCA78-300	7.25–7.75	27	1.2	0.5	13	23
JCA910-3000	9.0–9.5	25	1.3	0.5	13	23
JCA1112-3000	11.7–12.2	27	1.4	0.5	13	23
JCA1415-3001	14.4–15.4	35	1.6	1.0	14	24
JCA1819-3001	18.1–18.6	25	2.0	0.5	10	20
JCA2021-3001	20.2–21.2	25	2.5	0.5	10	20

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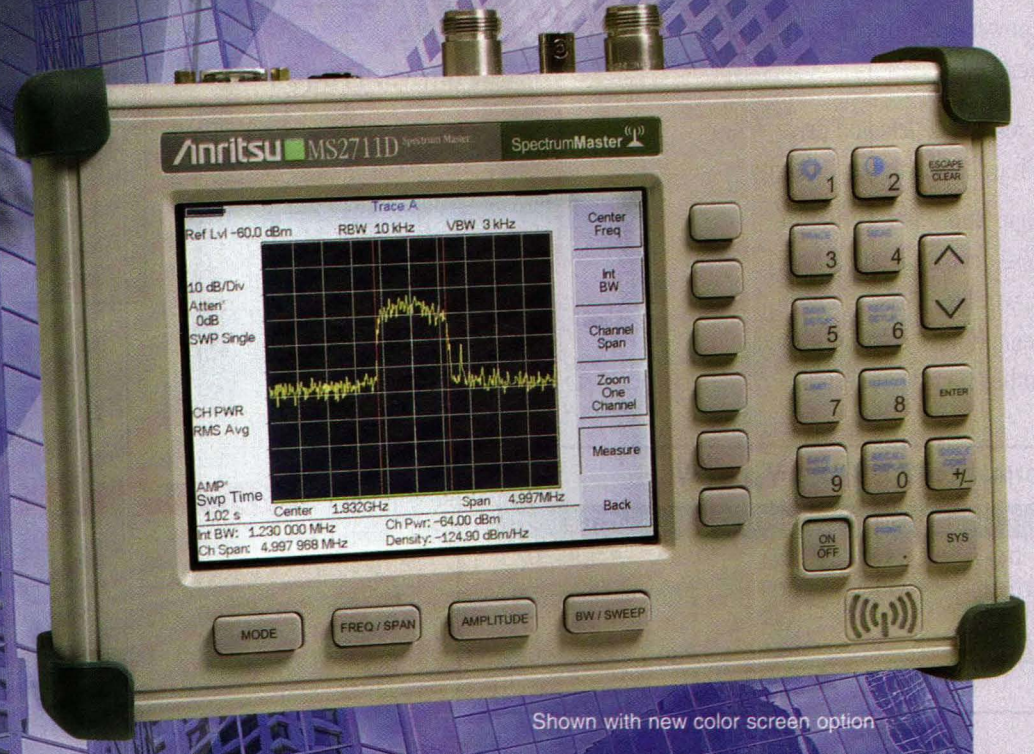

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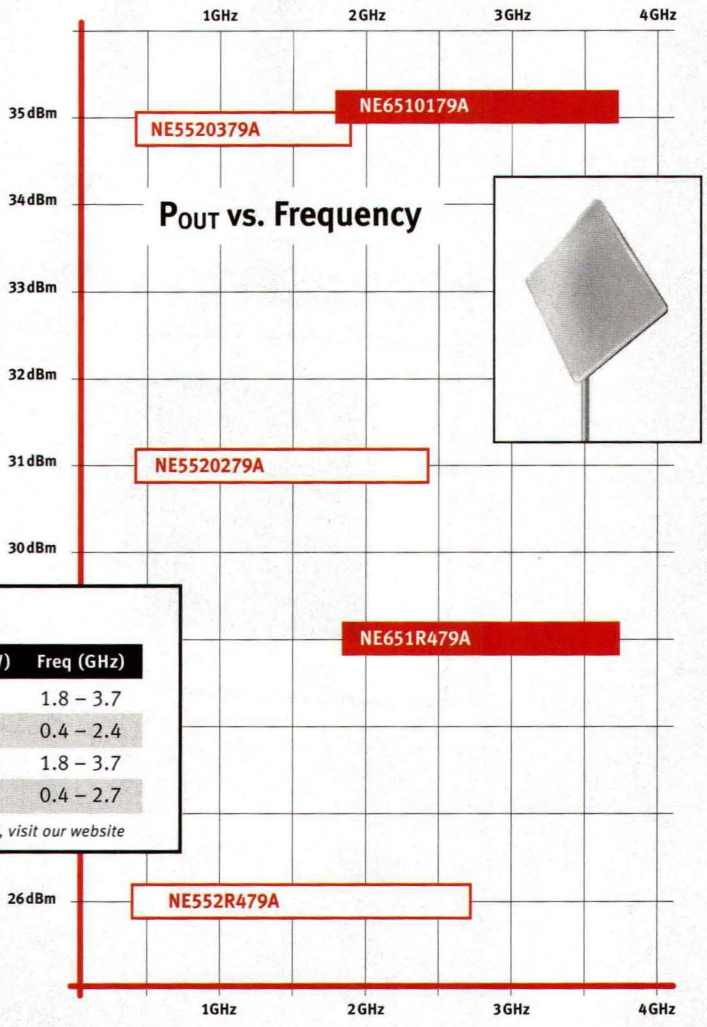

Typical Performance @ 2.3GHz, V_{DD} = 5V

Part Number	Type	P_{1dB} (dBm)	G_L (dB)	R_{TH} (°C/W)	Freq (GHz)
NE6510179A	GaAs	35	11	5	1.8 – 3.7
NE5520279A	LDMOS	31	10	7	0.4 – 2.4
NE651R479A	GaAs	29	12	12	1.8 – 3.7
NE552R479A	LDMOS	26	11	10	0.4 – 2.7

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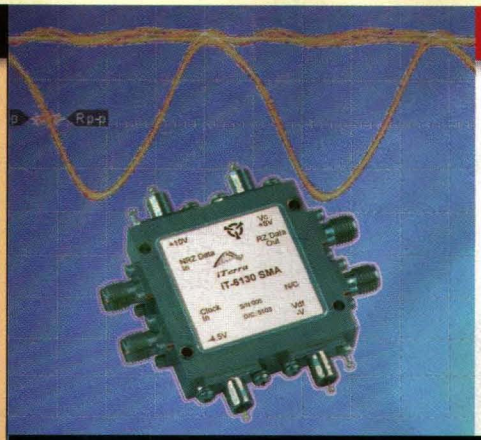
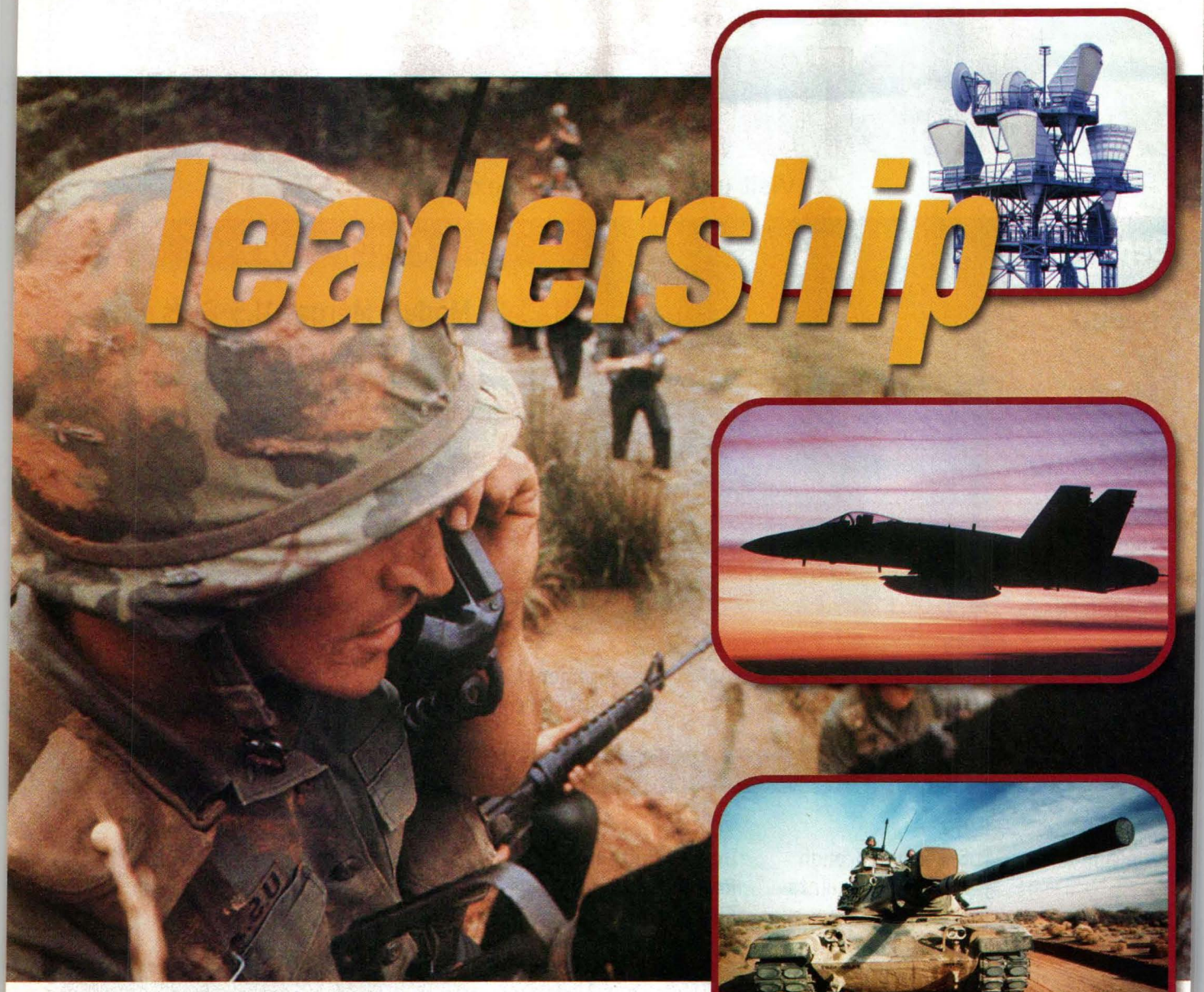


COVER STORY

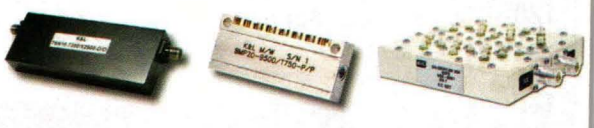
News

Design

Product Technology



Vision inspires Innovation.
Innovation defines
Leadership.



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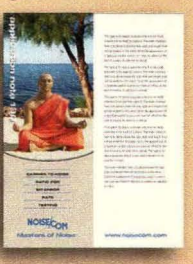

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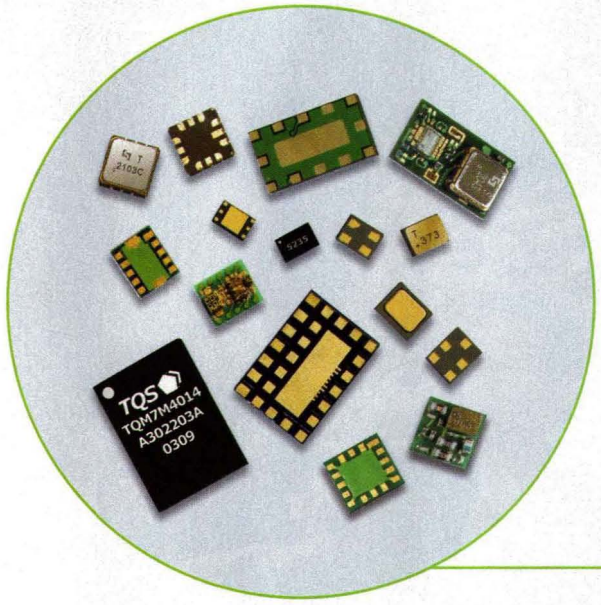

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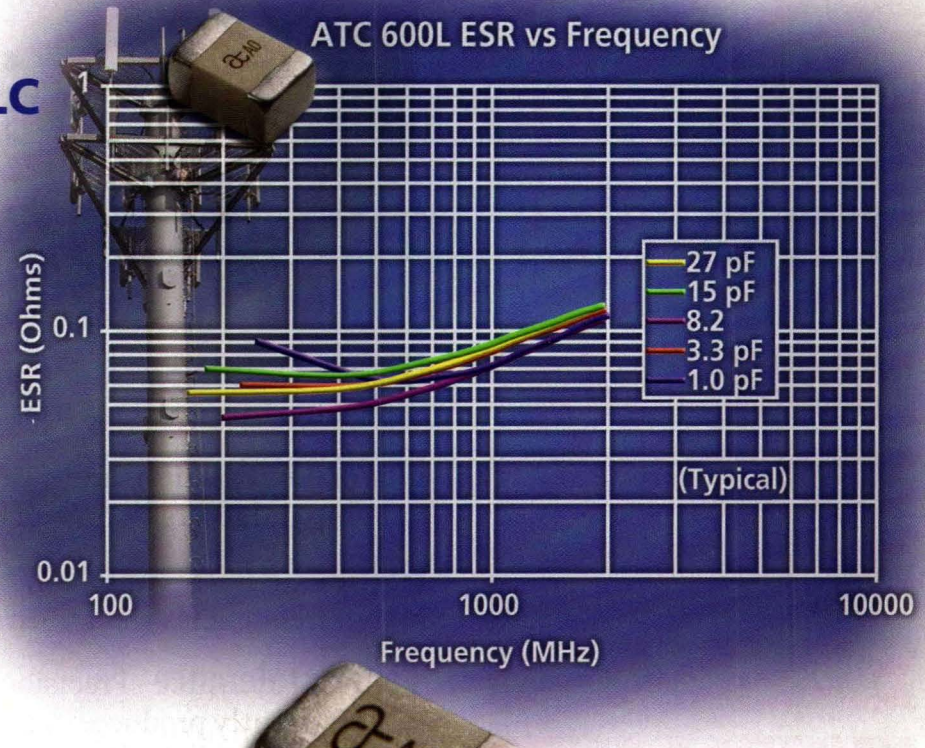

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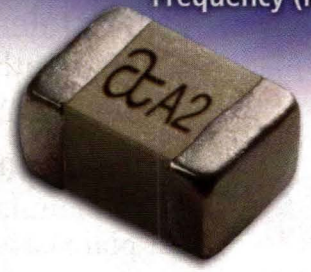

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Website Correction

▶▶PLEASE NOTE THAT the website mentioned in the author contact section of my article ("Evaluate The Performance Of Amplifying Predistorters," January, p. 84), www.rbtech-inc.com, is not my website. I have nothing to do with that company, and have never been associated with them in the past. The inclusion of that website in the author contact section of my article was a mistake made by *Microwaves & RF* while editing. RB Technology is my consulting firm located in Milpitas, CA, and registered in San Jose, CA.

Somnath Mukherjee
RB Technology
Milpitas, CA

Editor's Note: We apologize to the authors of the article, Somnath Mukherjee and Ralph Inducta, for the incorrect website listing. Microwaves & RF *regrets the error, and we apologize for any confusion or inconvenience that it may have caused.*

The DSO Arena

▶▶AS A PERSON with more than a passing familiarity with Tek, LeCroy, and Agilent products in the DSO arena, I'd like to make a short comment on a spec that is not often talked about: re-trigger latency/acquisitions per second.

While there are many specs of interest to potential purchasers of scopes, which all have different weighting in product selection, one that really stands out in the Tek product is the ability to capture trigger events that are closely spaced in time. As Jack Browne noted in his article (Special Report, "DSOs Track Elusive High-Speed Waveforms," January 2004, p. 106), there is the potential to capture 400k events per second in certain models. This can make a significant difference in acquiring some signals, as other scopes are limited to something like 3k to 10k acquisitions per second. When looking for an elusive glitch, this can make a huge difference in the time required to capture a random or pseudorandom event, and certainly makes a large difference in looking at signal statistics.

Barry Rowland
Vancouver, BC, Canada

PLEASE COMMENT

Microwaves & RF welcomes mail from its readers. Letters must include the writer's name and address. The magazine reserves the right to edit letters appearing in "Feedback." Address letters to:

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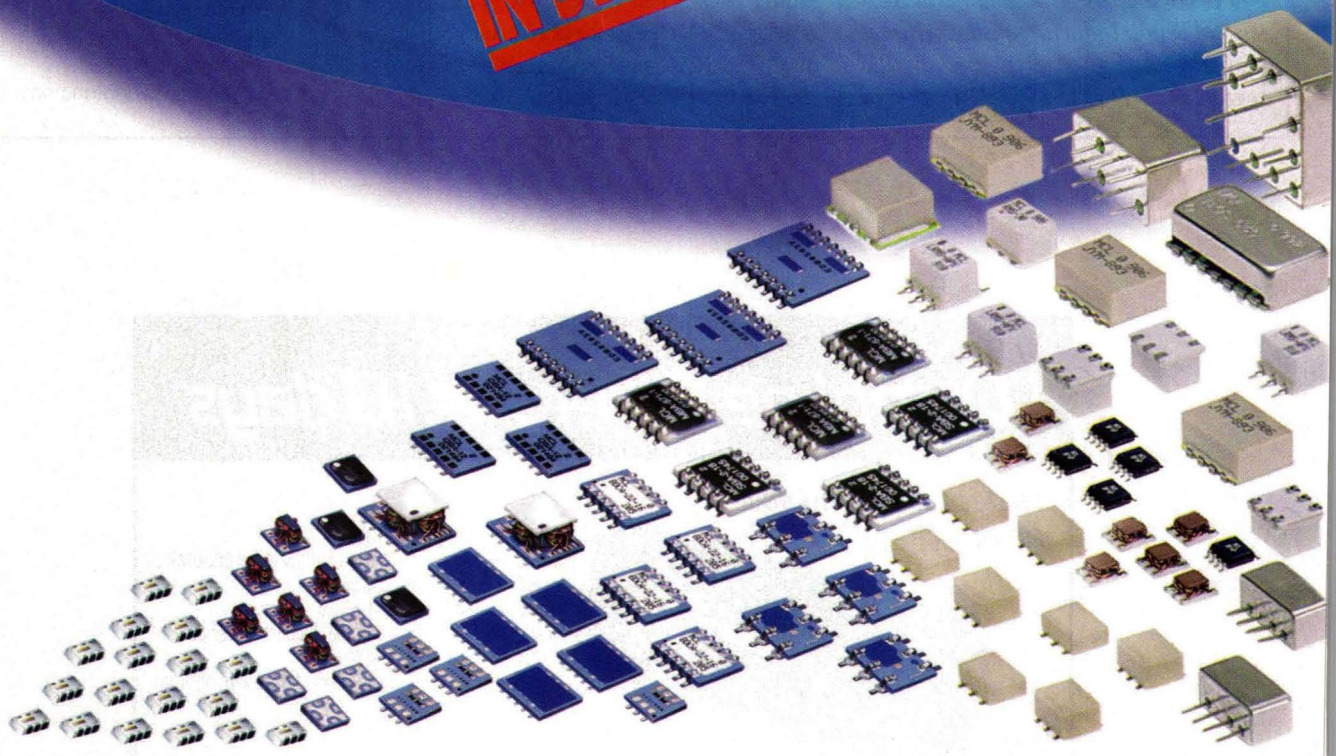

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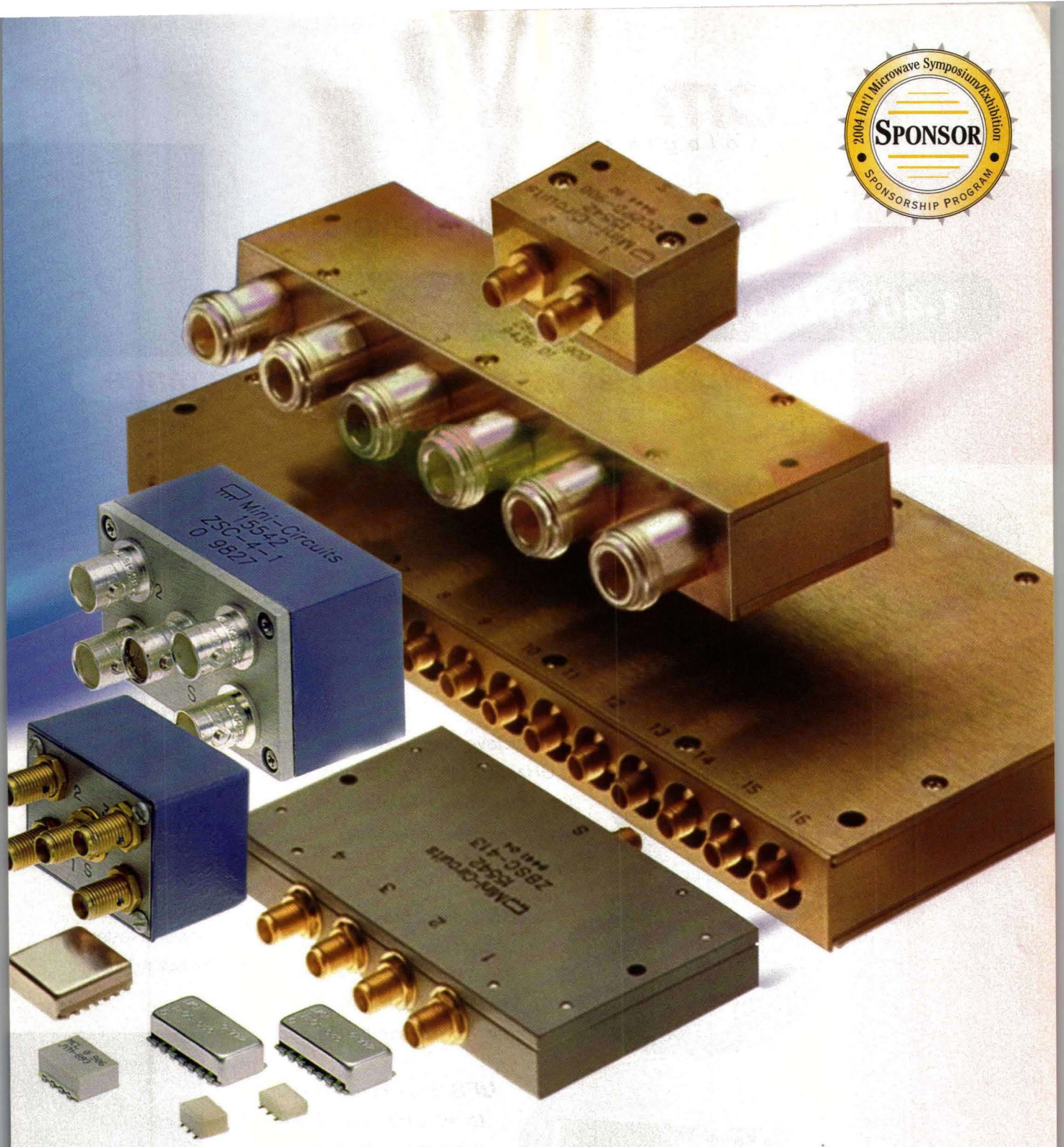

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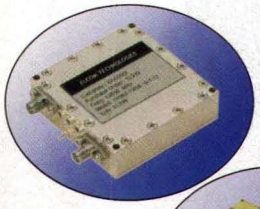
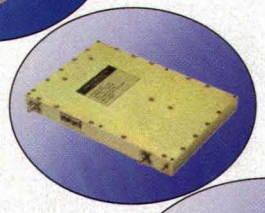


SPDRO *1-26 GHz. Phase locked in a single loop, 5-200 MHz reference, -130 dBc @ 100KHz. Offset 10 GHz, Temperature range from -55 to +85 celsius*



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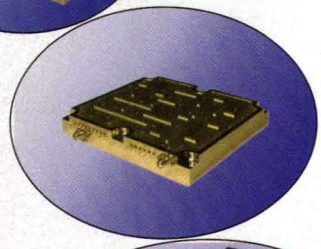
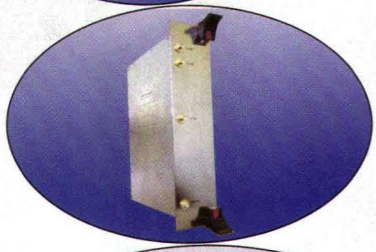


WMFS *Octave bandwidth indirect, up to 18 GHz, 1Hz step, 100 ms switching, Low phase noise, Ruggedized, Hermetically sealed package, Compact size 5"x5"*



IBS *Indirect broadbad synthesizer, .05 - 18 GHz, 10 - 250ms switching speed, VME/Rack configurations, ATE, Radar, Threat-simulation apps*

UFS *Broadband direct up to 40 GHz, 200 NSEC switching,-153 dBc phase noise @100 KHz offset from10 GHz signal*

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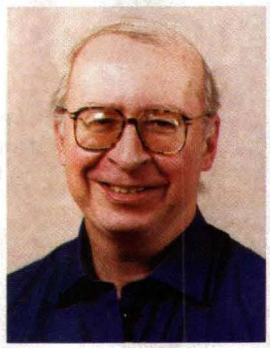

WIRELESS MARKETS have yielded little excitement in recent years, due largely to the slowdown in cellular infrastructure sales and the "commoditization" of products for wireless-local-area-network (WLAN) systems. Still, the curious came this past month (March 8-10) for the 12th running of the annual Wireless Systems Design Conference & Expo, held for the first time in San Diego, CA. Although the crowd was small, the technical program held their interest in a variety of different areas.

The San Diego conference featured multiple conferences within a conference.

The San Diego conference featured multiple conferences within a conference. In addition to the expected wireless technology/applications presentations, the event featured a small program devoted to military electronics and a first-time program sponsored by sister Penton publication, *Machine Design*, called the Industrial Wireless Applications Summit (IWAS). Each event featured its own Keynote Speaker, with Dr. Henry Samueli, co-founder and chairman of Broadcom Corp. (Irvine, CA) tackling the task for the Wireless Systems Design Conference. Broadcom, founded in 1991 and taken public in 1998, is one of the bright spots in a struggling wireless industry, having diversified into wired and wireless markets for semiconductors in a variety of applications, including cellular, cable, satellite communications, and WLANs.

In a talk entitled "Wireless in Everything: Life in a Fully Connected World," Dr. Samueli foretold a future where wireless technology could be used in almost all things electronic. He emphasized that this future would be made possible by low-cost wireless transceivers (including controllers) on a single chip.

On the military side, Dr. Ronald E. Reedy, founder, vice president, and chief technology officer (CTO) of Peregrine Semiconductor (San Diego, CA) offered advice on supplementing sagging wireless business with higher-margin military business. He detailed in particular an elegant Global Positioning System (GPS) receiver module for the military that brought the company the added benefits of scaled-down versions of the module for commercial sales.

The third of the trio of Keynote speakers. Robert Poor, CTO and co-founder of Ember Corp. (Boston, MA) spoke on the industrial side of wireless. He emphasized that a healthy wireless market is one that does not rely on just one business segment (cellular), but is diversified into such areas as medical and industrial applications.

A great deal of thanks and appreciation are owed to the Keynote speakers, workshop hosts, and presenters who not only made the journey to San Diego, but gave so generously of their time, as well as to those attending the 12th annual meeting. For those who missed it, next month will carry a highly condensed summary.

Jack Browne

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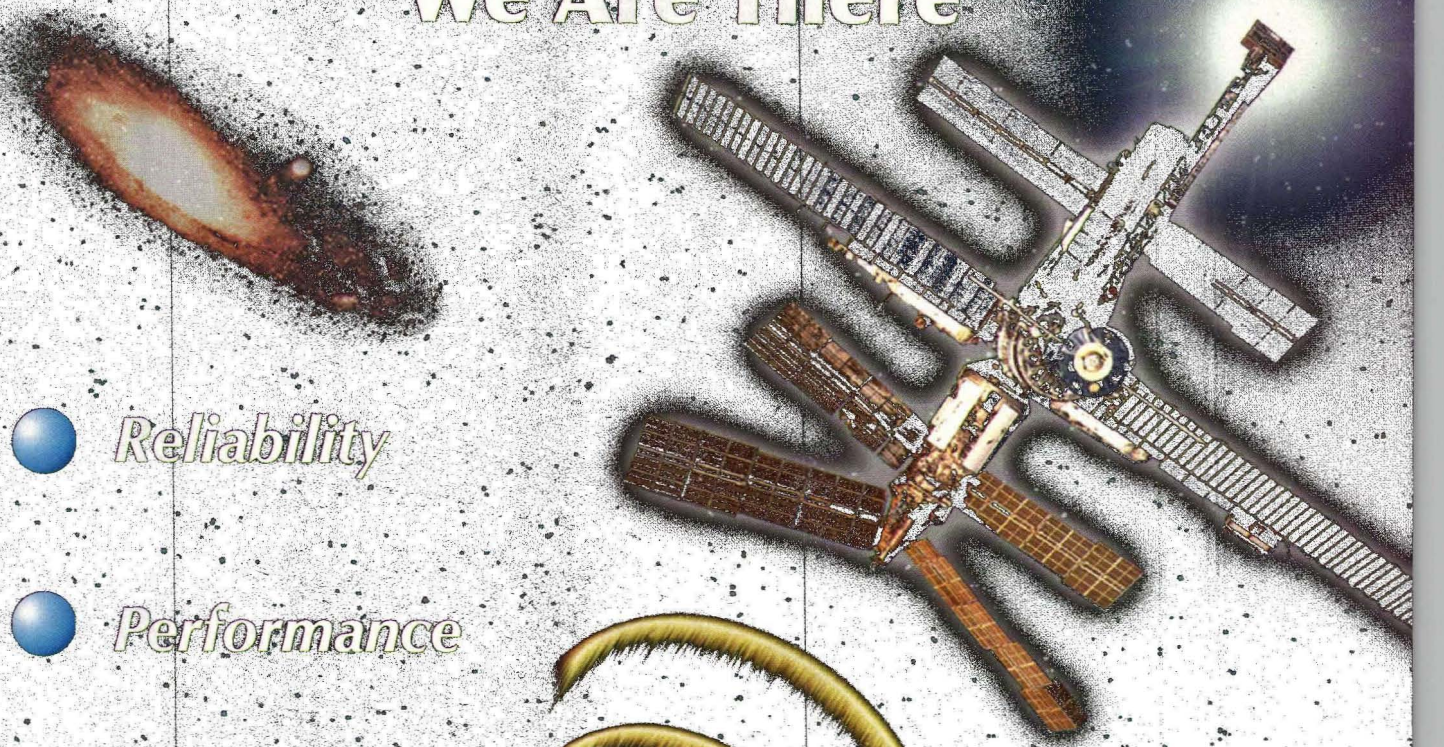

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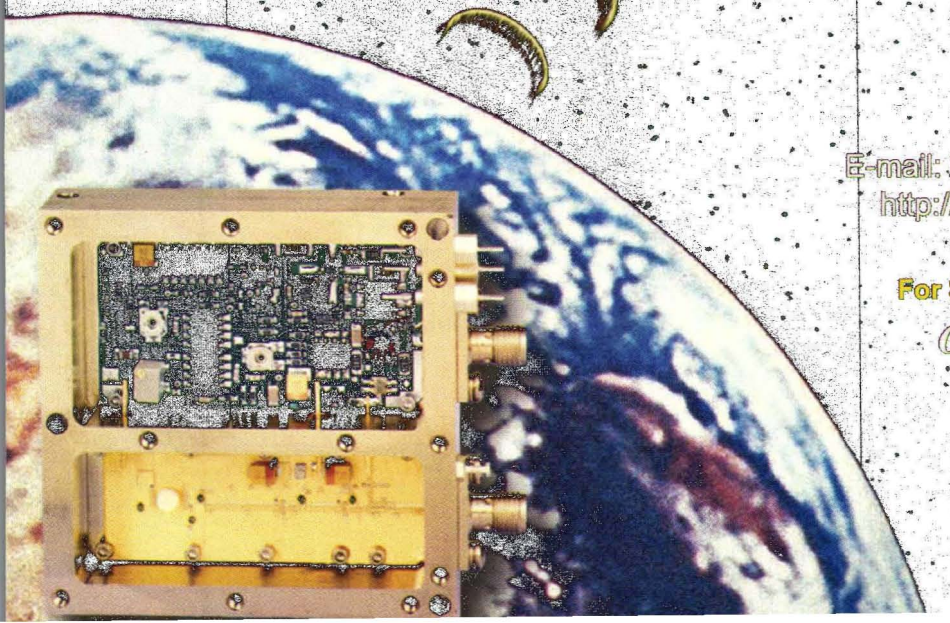

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OCTAVE BAND AMPLIFIERS								
JS2-00500100-045-5A	0.5 – 1	35	1	0.45	2:1	2:1	5	250
JS2-00500100-12-5A	0.5 – 1	35	1.2	1	2:1	2:1	5	250
JS2-01000200-045-5A	1 – 2	33	1	0.45	2:1	2:1	5	250
JS2-02000400-045-5A	2 – 4	28	1.2	0.45	2:1	2:1	5	175
JS2-04000800-08-0A	4 – 8	22	1.2	0.8	2:1	2:1	0	150
JS3-04000800-08-5A	4 – 8	30	1	0.8	2:1	2:1	5	175
JS3-04000800-15-5A	4 – 8	30	1	1.5	2:1	2:1	5	175
JS2-08001200-11-5A	8 – 12	15	1	1.1	2:1	2:1	5	150
JS3-08001200-11-5A	8 – 12	25	1	1.1	2:1	2:1	5	175
JS3-08001200-15-5A	8 – 12	25	1	1.5	2:1	2:1	5	175
JS3-12001800-16-5A	12 – 18	23	1	1.6	2:1	2:1	5	175
JS4-12001800-145-5A	12 – 18	30	1	1.45	2:1	2:1	5	200
JS4-12001800-30-5A	12 – 18	30	1	3	2:1	2:1	5	200
JS2-18002600-20-5A	18 – 26	14	2	2	2.5:1	2.5:1	5	100
JS2-18002600-30-5A	18 – 26	14	2	3	2.5:1	2.5:1	5	100
JS3-18002600-20-5A	18 – 26	22	1.8	2	2.5:1	2.5:1	5	175
JS3-18002600-30-5A	18 – 26	22	1.8	3	2.5:1	2.5:1	5	175
JS4-18002600-19-5A	18 – 26	33	1.5	1.9	2:1	2:1	5	200
JS4-18002600-26-5A	18 – 26	33	1.5	2.6	2:1	2:1	5	200
JS2-26004000-35-5A	26 – 40	10	2	3.5	2.5:1	2.5:1	5	100
JS2-26004000-45-5A	26 – 40	10	2	4.5	2.5:1	2.5:1	5	100
JS3-26004000-35-5A	26 – 40	18	2.5	3.5	2.5:1	2.5:1	5	175
JS3-26004000-45-5A	26 – 40	18	2.5	4.5	2.5:1	2.5:1	5	175
JS4-26004000-40-5A	26 – 40	23	2.5	4	2:1	2:1	5	200
JS4-40006000-65-0A	40 – 60	15	3	6.5	2.75:1	2.75:1	0	175
MULTIOCTAVE BAND AMPLIFIERS								
JS2-00500200-07-5A	0.5 – 2	32	1	0.7	2:1	2:1	5	295
JS2-00500200-15-5A	0.5 – 2	32	1	1.5	2:1	2:1	5	295
JS2-01000400-08-5A	1 – 4	27	1	0.8	2:1	2:1	5	200
JS2-01000400-20-5A	1 – 4	27	1	2	2:1	2:1	5	200
JS2-02000600-08-5A	2 – 6	22	1	0.8	2:1	2:1	5	125
JS2-02000600-20-5A	2 – 6	22	1	2	2:1	2:1	5	125
JS2-02000800-08-0A	2 – 8	22	1.25	0.8	2:1	2:1	0	125
JS2-02000800-20-0A	2 – 8	18	1.25	2	2:1	2:1	0	125
JS3-02001800-25-5A	2 – 18	23	1.8	2.5	2.5:1	2.5:1	5	150
JS3-02001800-50-5A	2 – 18	23	1.8	5	2.5:1	2.5:1	5	150
JS4-02001800-22-5A	2 – 18	30	2	2.2	2.5:1	2.5:1	5	200
JS4-02001800-50-5A	2 – 18	30	2	5	2.5:1	2.5:1	5	200
JS3-02002600-33-5A	2 – 26	21	2.5	3.3	2.5:1	2.5:1	5	150
JS3-02002600-40-5A	2 – 26	21	2.5	4	2.5:1	2.5:1	5	150
JS3-06001800-16-5A	6 – 18	23	1.8	1.6	2:1	2:1	5	125
JS3-06001800-30-5A	6 – 18	23	1.8	3	2:1	2:1	5	125
JS4-06001800-145-5A	6 – 18	31	2	1.45	2:1	2:1	5	200
JS4-06001800-30-5A	6 – 18	31	2	3	2:1	2:1	5	200

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MULTIOCTAVE BAND AMPLIFIERS (continued)								
JS3-08001800-16-5A	8 – 18	24	1.5	1.6	2:1	2:1	5	150
JS3-08001800-30-5A	8 – 18	24	1.5	3	2:1	2:1	5	150
JS4-08001800-145-5A	8 – 18	32	2	1.45	2:1	2:1	5	200
JS4-08001800-30-5A	8 – 18	32	2	3	2:1	2:1	5	200
JS3-12002600-25-5A	12 – 26	22	2.5	2.5	2.2:1	2.2:1	5	150
JS3-12002600-35-5A	12 – 26	22	2.5	3.5	2.2:1	2.2:1	5	150
JS4-12002600-22-5A	12 – 26	32	2.2	2.2	2:1	2:1	5	200
JS4-12002600-35-5A	12 – 26	32	2.2	3.5	2:1	2:1	5	200
JS3-18004000-38-5A	18 – 40	16	2.5	3.8	2.5:1	2.5:1	5	150
JS3-18004000-50-5A	18 – 40	16	2.5	5	2.5:1	2.5:1	5	150
JS4-18004000-30-5A	18 – 40	23	2.5	3	2.5:1	2.5:1	5	200
JS4-18004000-50-5A	18 – 40	23	2.5	5	2.5:1	2.5:1	5	200
ULTRAWIDE BAND AMPLIFIERS								
JS2-00100200-07-5A	0.1 – 2	32	1	0.7	2:1	2:1	5	295
JS2-00100200-15-5A	0.1 – 2	32	1	1.5	2:1	2:1	5	295
JS2-00100400-08-5A	0.1 – 4	27	1	0.8	2:1	2:1	5	200
JS2-00100400-12-5A	0.1 – 4	27	1	1.2	2:1	2:1	5	200
JS2-00100600-10-3A	0.1 – 6	23	1.5	1	2:1	2:1	3	175
JS2-00100600-20-3A	0.1 – 6	23	1.5	2	2:1	2:1	3	175
JS2-00100800-13-0A	0.1 – 8	20	1.5	1.3	2:1	2:1	0	175
JS2-00100800-25-0A	0.1 – 8	20	1.5	2.5	2:1	2:1	0	175
JS3-00101000-20-5A	0.1 – 10	23	1.5	2.0	2.5:1	2:1	5	150
JS3-00101000-35-5A	0.1 – 10	23	1.5	3.5	2.5:1	2:1	5	150
JS3-00101200-21-5A	0.1 – 12	23	1.5	2.1	2.5:1	2:1	5	150
JS3-00101200-35-5A	0.1 – 12	23	1.5	3.5	2.5:1	2:1	5	150
JS3-00101800-24-5A	0.1 – 18	23	1.8	2.4	2.5:1	2.2:1	5	150
JS3-00101800-40-5A	0.1 – 18	23	1.8	4	2.5:1	2.2:1	5	150
JS4-00101800-23-5A	0.1 – 18	29	1.8	2.3	2.5:1	2.2:1	5	200
JS4-00101800-40-5A	0.1 – 18	29	1.8	4	2.5:1	2.2:1	5	200
JS4-00102000-25-5A	0.1 – 20	28	1.8	2.5	2.5:1	2.5:1	5	200
JS4-00102000-35-5A	0.1 – 20	28	1.8	3.5	2.5:1	2.5:1	5	200
JS3-00102600-33-5A	0.1 – 26	20	2.5	3.3	2.5:1	2.5:1	5	150
JS3-00102600-42-5A	0.1 – 26	20	2.5	4.2	2.5:1	2.5:1	5	150
JS4-00102600-28-5A	0.1 – 26	27	2.5	2.8	2.5:1	2.5:1	5	200
JS4-00102600-50-5A	0.1 – 26	27	2.5	5	2.5:1	2.5:1	5	200
JS4-00104000-65-5A	0.1 – 40	14	4.5	6.5	2.75:1	2.75:1	5	200
JS4-00104000-85-5A	0.1 – 40	14	4.5	8.5	2.75:1	2.75:1	5	200

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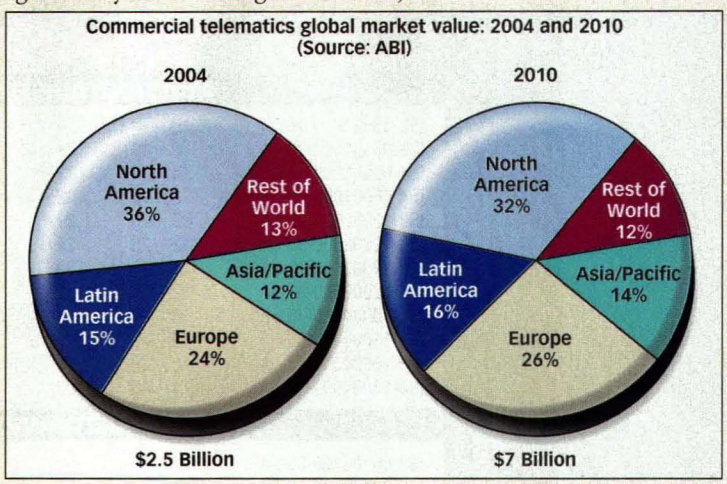
the front end

News items from the communications arena.

Wireless Carriers Are Now Joining The Commercial Telematics Bandwagon

OYSTER BAY, NY—A new crop of sophisticated, location-enabled wireless handsets is helping wireless carriers to integrate themselves deeper into the commercial telematics value chain according to "Fleet Management Systems Intelligence Service," a report from technology research firm ABI.

Because of relatively high costs for embedded telematics hardware, the barriers of entry are still too high for most of the smaller fleet operators. As a result, many commercial telematics vendors have had difficulty reaching this substantial, yet virtually untapped market with a potential of $7 billion by 2010 **(see figure)**.



"Wireless carriers are uniquely positioned to address a specific niche in commercial telematics," states Frank Viquez, ABI's director of automotive research. "By fully leveraging its pre-existing relationships with hardware and software partners, a wireless carrier can tap into its subscriber base and target particular commercial customers who previously wouldn't have been able to consider a telematics solution."

ABI identifies Nextel Communications as the first and best positioned North American carrier to offer a handset-based commercial telematics solution. Nextel's 13 million subscribers are predominantly business users who spend nearly a third more on their phone bills than the average consumer subscriber.

GSM Is On Track To Connect Its Billionth Customer

LONDON, ENGLAND—Global System for Mobile Communications (GSM) had 970 million users at the end of December 2003 and an average of 15 million new users a month throughout 2003. The GSM Association, the global trade association representing GSM operators worldwide, expected the billionth GSM user to be connected during the first quarter of 2004. The approach of this milestone was a major theme at the 3GSM World Congress, the annual gathering of the global GSM community, which took place in Cannes, France from February 23 to 26.

During 2003, more than 180 million new users joined the global GSM community. These new users represented 80 percent of the 227 million new digital mobile-phone users connected worldwide.

Rob Conway, chief executive of the GSM Association and member of its board, comments, "By taking 80 percent of its growth in 2003, GSM outperformed the market and reinforced its position as the global standard for mobile communications."

For the second year running, GSM added as many customers during the year as the second-most-successful mobile technology (CDMA) had in total at the year's end.

In absolute terms, members of the Association's Asia Pacific regional interest group drove GSM's growth during the year, adding more than 70 million new users. China maintained its position as the world's largest mobile market: China's two GSM operators added 42.8 million. Europe contributed 42 million new users.

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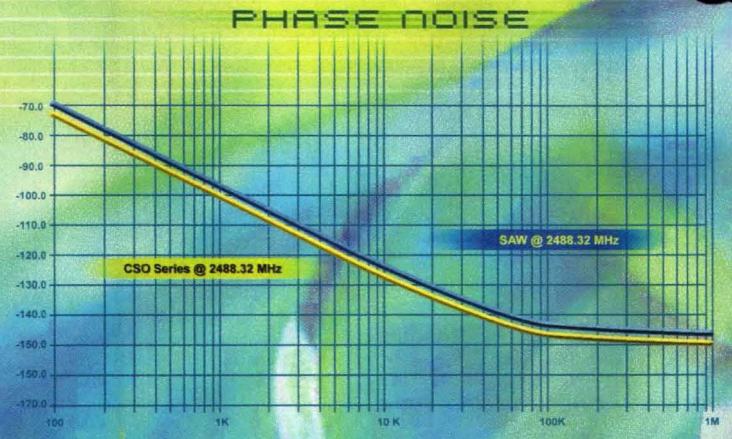

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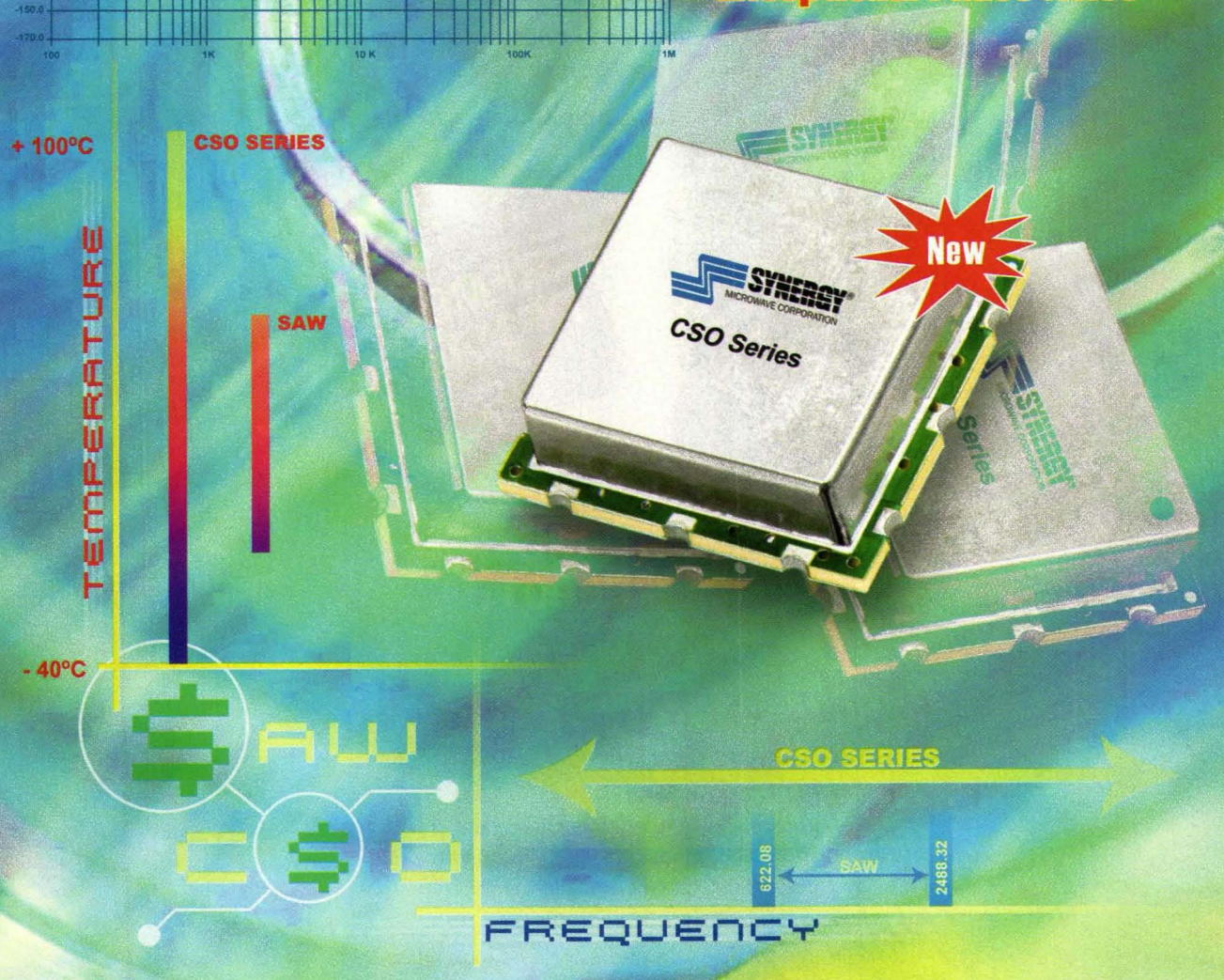

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- Digital and analog I/Q baseband interfaces

Agere Systems Earns Device Wins in 3G Systems

ALLENTOWN, PA—Agere Systems, one of the world's leading suppliers of semiconductors for wireless systems, announced two major business transactions with builders of third-generation (3G) wireless systems. The company disclosed that is has been supplying high-power laterally diffused metal-oxide-semiconductor (LDMOS) transistors to NEC (Tokyo, Japan) for use in third-generation (3G) wideband code-division-multiple-access (WCDMA) wireless base-station equipment. "The use of Agere transistors in our base stations will accelerate our company's 3G wireless equipment deployment during the next few years," comments Dr. Nobuhiro Endo, general manager of NEC's Mobile and Wireless Division, "Power efficiency is extremely important in a Wideband CDMA base-station design, and Agere's transistors delivered the performance required," he adds. Agere also equips NEC with traffic management chips for use in NEC's high-speed router family, and General Packet Radio Service (GPRS) wireless data chips for NEC cellular phones.

In a separate announcement, Agere unveiled information about supplying high-power RF transistors to Korean telecommunications firm Sewon Teletech, Inc. for use in 3G wireless infrastructure equipment. Sewon is the largest supplier of power amplifiers to wireless and repeater original equipment manufacturers in Korea. Sewon uses Agere's RF LDMOS transistor-based amplifiers in Korea's CDMA-based personal-communication-services (PCS) wireless networks. "Agere's technology has proven to be reliable and technically sound for our power amplifiers," says Dr. Chul Dong Kim, Sewon's CEO. Sewon's amplifiers have been in service in the market for more than six months with no failures. According to Carlos Garcia, vice president and general manager of Agere's Analog Products Division, "Sewon is one of Agere's key Korean customers. Both companies are working together to accelerate Korea's installation of next-generation wireless equipment, which already ranks among the world's most advanced."

For additional information about Agere Systems and the products and support that they offer, see the company's website at www.agere.com.

> ***"Adding an RFID tag to something creates so much added value, security, and utility."***

SkyeTek's RFID Technology Is Used At The Golden Globe Awards

HOLLYWOOD, CA AND BOULDER, CO—Skyetek, Inc., RFID reader technology provider, and Security Solutions, Inc. have revealed how RFID was used at the 2004 Golden Globe Awards.

Each invitation and event credential used at this year's Golden Globes contained an RFID tag inconspicuously embedded inside. This simple measure completely eliminated the risk of counterfeiting and increased security overall, while making movement and access faster and easier. The credentials, when read with the Skyetek reader attached to a PC or Tablet PC at key checkpoints, would display the photo and access right for instant visual verification. The invited celebrity guests presented their invitations to security personnel, who read the invitation with Skyetek portable RFID readers to confirm the authentication of the invitation and enroll the guest as "arrived" on a host server.

"This is a perfect example of how adding an RFID tag to something, in this case to an invitation, creates so much added value, security, and utility. And portable low-power RFID technology from Skyetek continues enabling innovative RFID such as this all across the 'golden globe,'" says Skyetek president, Jonathan Bein, Ph.D. Using the Skyetek RFID technology, Security Solutions was able to implement the application in a rapid time frame. Tim DeWeese, Security Solutions' president, mentioned that the system employed an integration of several technologies to provide the solution required by the Golden Globes. Security measures included: a bank of photo ID credentialing workstations; networks both wired and wireless to provide "real time" updates to the system; and RFID tags and readers to provide access control credentials and document authentication all tied to one system.

RFID (radio-frequency identification) first appeared in tracking and access applications during the 1980s. RFID wireless systems allow for non-contact reading and are effective in manufacturing and other hostile environments where bar-code labels cannot survive. RFID systems have tags and readers. Tags are microchip-based radios with built-in antennas, which have already shrunk to the size of postage stamps. They listen for a radio query from a reader and respond by transmitting their unique ID code. RFID is established in a wide range of markets, including retail, health care, and livestock identification.

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S2W2	S2W5	N2W5	2	±0.40
S3W2	S3W5	N3W5	3	±0.40
S4W2	S4W5	N4W5	4	±0.40
S5W2	S5W5	N5W5	5	±0.40
S6W2	S6W5	N6W5	6	±0.40
S7W2	S7W5	N7W5	7	±0.60
S8W2	S8W5	N8W5	8	±0.60
S9W2	S9W5	N9W5	9	±0.60
S10W2	S10W5	N10W5	10	±0.60
S12W2	S12W5	N12W5	12	±0.60
S15W2	S15W5	N15W5	15	±0.60
S20W2	S20W5	N20W5	20	±0.60
S30W2	S30W5	N30W5	30	±0.85
S40W2	S40W5	N40W5	40	±0.85

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As WCDMA Networks Grow, WCDMA's Role Becomes Defined

KINGS PARK, NY—Although off to slower start than originally expected, WCDMA is showing regained promise, according to a new study from Visant Strategies. Current WCDMA deployments have no more than small parts of nations covered. Some very atypical rollouts exist, but the need for voice capacity and a high-end data niche will allow WCDMA carriers to do well, the study finds.

According to "WCDMA: Promise of Success 2004," close to 80,000 WCDMA base stations are live as of the first quarter of 2004, and shipments of WCDMA infrastructure will continue to rise, driven by coverage requirements in Europe and market developments in Japan.

"WCDMA is expected to enjoy greater than 60-percent penetration in both Japan and Western Europe in 2009," comments report author Andy Fuertes, a Visant senior analyst. "WCDMA will play a lesser role outside of these areas due to limited demand for enhancements in voice capacity in the short term, although it is reemerging as an option in North America due to increasing competition from other air interfaces."

"WCDMA is expected to enjoy greater than 60% penetration in both Japan and Western Europe in 2009."

One critical assumption made in the study is the ability for the wireless industry to produce a capable and relatively inexpensive WCDMA phone during the 2005 to 2007 timeframe, allowing unknowing 2G and 2.5G users to be ported over to WCDMA seamlessly. Gradual improvements in content, applications, input/output, and screen technology will drive broad interesting mobile Internet services but handset cost and requirements for voice capacity are the triggers which will lead carriers to migrate their existing user base to WCDMA.

Also, although current ARPU to WCDMA services is 30 percent higher than the average ARPU derived from 2G users, such increases are unsustainable within mass-market acceptance. Today, the potential WCDMA user is defined by price-conscious, prepaid customers in most of the markets in which technology has been deployed.

The study provides global and regional annual WCDMA chip-set shipments and revenues for 2003 to 2009. It also details global and regional annual WCDMA subscribers and WCDMA handset shipments for 2003 to 2009. Also detailed are the impact of competitors to WCDMA, such as 802.20, CDMA2000, EDGE, and TD-SCDMA.

Wireless And Microwave Confab Will Be Held In Clearwater

CLEARWATER, FL—The Wireless and Microwave Technology Conference 2004 will be held at the Marriot Sand Key in Clearwater, FL on April 15 and 16, 2004. The conference's technical program includes oral presentations, poster presentations, and tutorials. According to the conference's General Chair, Dr. Barry S. Perlman, this is the first time that the meeting has been open for general participation.

In his letter announcing the conference, Dr. Perlman wrote: "In the past, a similar but abbreviated one-day meeting was organized by the local IEEE Chapter and held on a Saturday just off campus. While those meetings were very successful, they were not generally publicized broadly and most participants were from the local Tampa Bay area. With the rapid growth of wireless and related microwave/RF technologies, very hot topics these days, the time has come to get the word out and publicize the great work going on in the WAMI Lab at USF (University of South Florida) and invite others with similar activities and interest to submit papers on complementary work and join together to experience the wonderful and exciting progress in this area."

For registration information and the conference agenda, see the conference's website at www.wami2004.org.

Kudos

NORWOOD, MA—Bill Gates has given the first public demonstration of Microsoft® Windows® Media Center Extender Technology—a powerful media adapter that will enable consumers to access their favorite digital entertainment that resides on their Windows XP Media Center Edition PC, from any room in the house regardless of where the PC is located. Analog Devices, Inc. has announced that its Blackfin® Processor manages the Windows Media audio/video decode processing in the new Media Center Extender set-top products. Having been selected by Microsoft, ADI's Blackfin Processor is part of the standard design that original equipment manufacturers (OEMs) and original design manufacturers (ODMs) will use to bring Windows Media Center Extender Technology to market in hardware designs that include set-top boxes and Media Center-ready televisions. MRF

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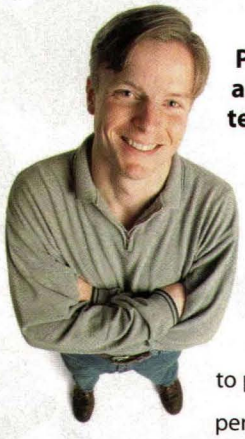

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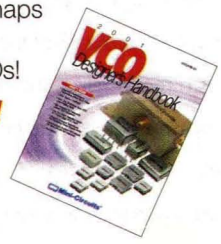

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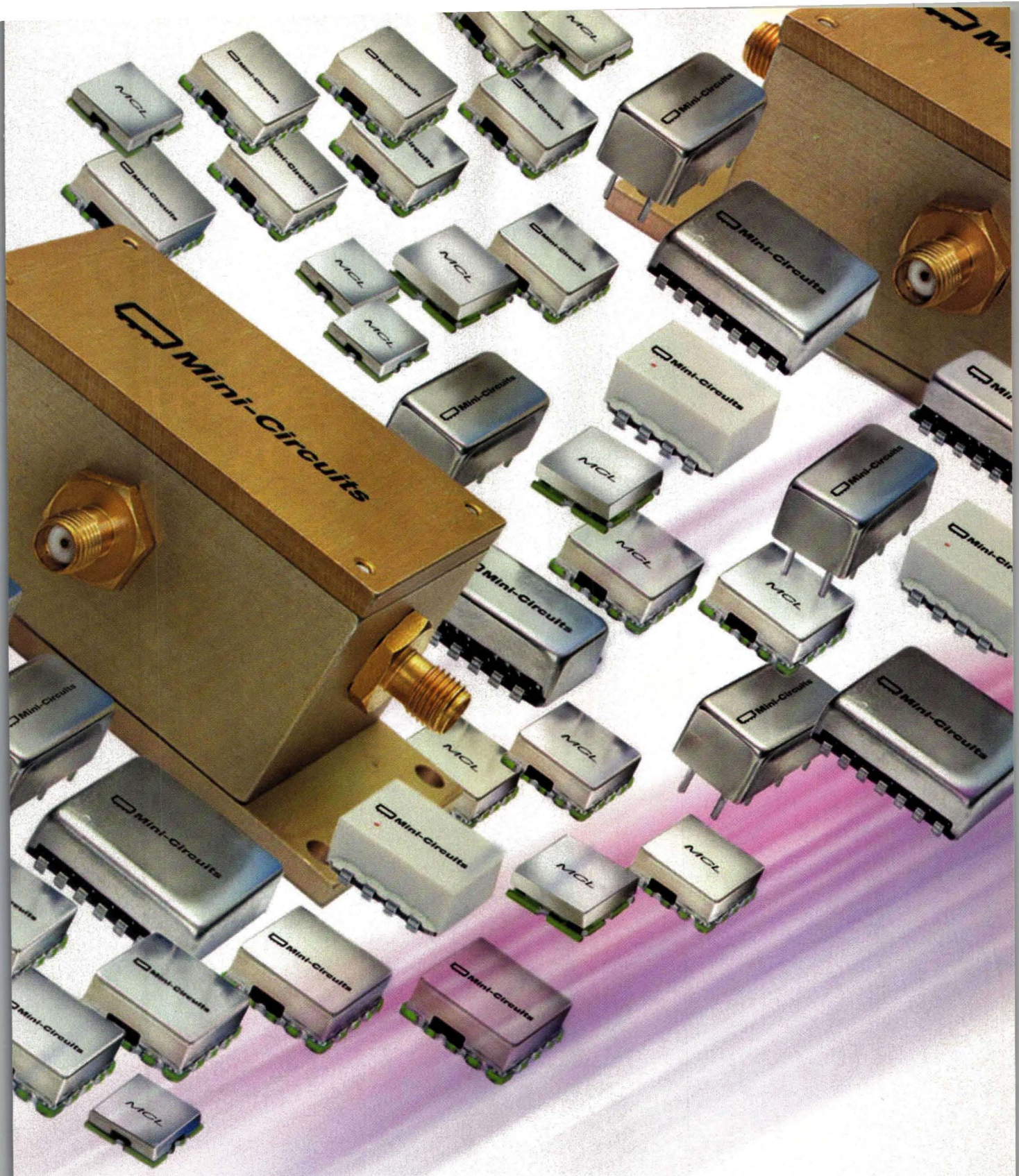

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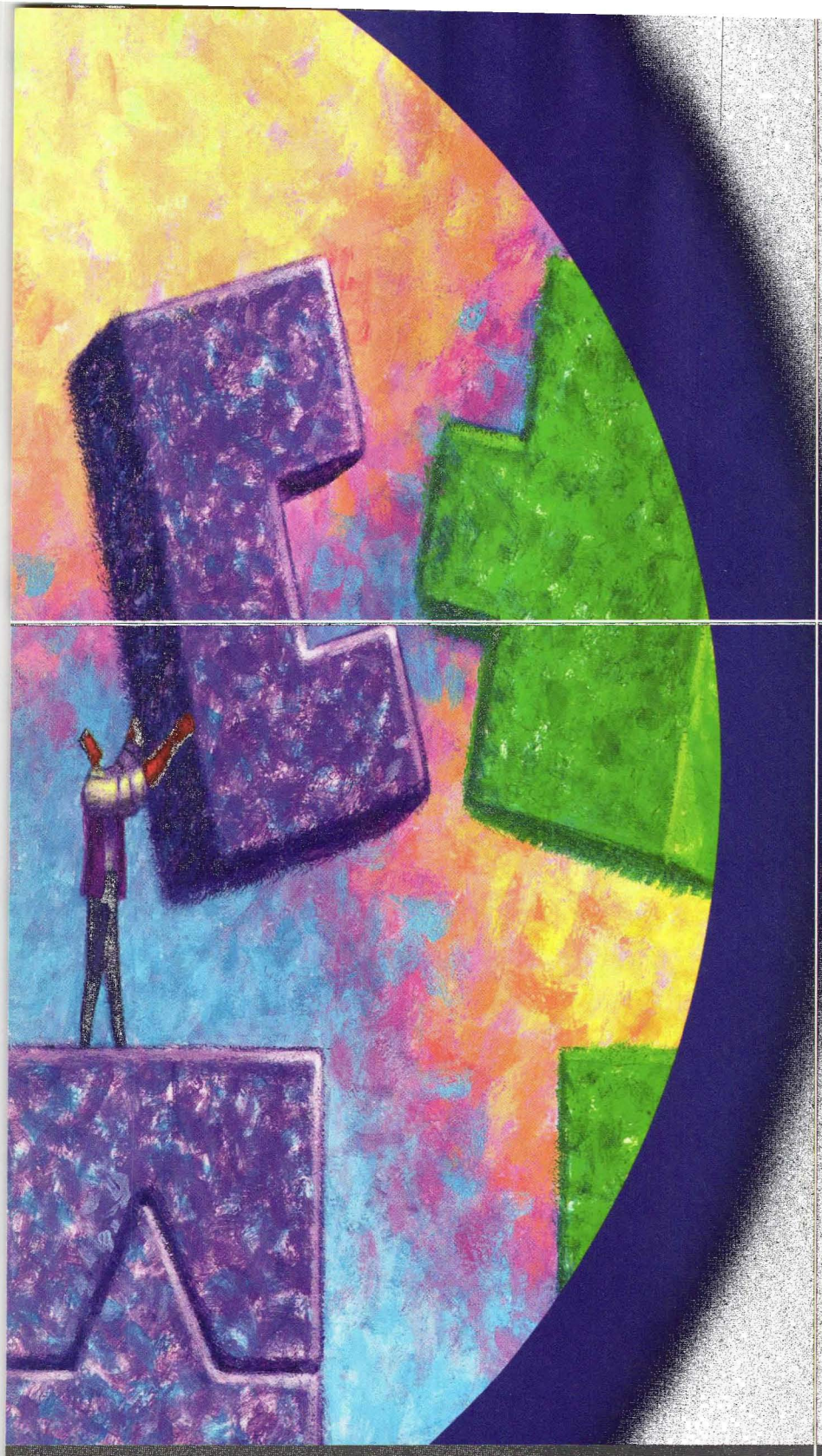

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NEWS

ADCs Clear Way To Digital Receivers

The ever-improving balance of bandwidth, resolution, and power consumption are making modern analog-to-digital converters an essential component in high-frequency systems.

Analog signals once dominated high-frequency designs, such as receivers. But with the growing availability of high-speed analog-to-digital converters (ADCs), with impressive capabilities in translating analog signals to the digital realm, an increasing amount of signal processing is being done by digital hardware. ADCs are currently available from a wide range of suppliers, in numerous formats include as chips, as plug-in circuit cards, and as printed-circuit-board (PCB) assemblies.

JACK BROWNE
Publisher/Editor

DAWN PRIOR
Editorial Assistant

ADCs can be readily differentiated by a handful of basic performance specifications, including bit resolution, input bandwidth, sampling rate, bit linearity, power consumption, noise performance, and output types. In operation, ADCs generate a digital code for a discrete value of input voltage. The number of codes is simply 2^N where N = the bit resolution of the ADC. An 8-b device, for example, uses 2^8 or 256 different digital codes to represent an analog waveform. Although high-performance digital-audio applications have standardized on the use of 24-b converters at sampling rates as high as 192 kSamples/s (kHz), RF and microwave applications generally rely on ADCs with anywhere from 6 to 14 b resolution. (For those wishing a thorough introduction to ADCs, the 69-page "ABCs of ADCs," written by Nicholas Gray, Staff Applications Engineer for the Data Conversion Systems of National Semiconductor, is available for free download from National's website at www.national.com.)

The number, N, of digital bits is directly related to the ADC's signal-to-noise ratio (SNR) and spurious-free dynamic range (SFDR). For an ADC, the dynamic range is the ratio of the largest resolvable signal to the smallest resolvable signal. For an ideal ADC it is 6.02N, or about 48 dB for a 8-b ADC. Thus, while it is desirable to have as many bits of resolution for a given application, a general trade-off for ADCs is decreasing bits for increasing bandwidth. Several suppliers, for example, offer ADCs with 1 GHz or greater input bandwidths, although the bit resolution is generally 8 b or less. Nyquist theory is a reliable guide when specifying an ADC for an application. According to Nyquist's criteria, an analog signal with a given bandwidth must be sampled with a sampling rate of at least two times that bandwidth to avoid loss of information. If sampling occurs at a rate of less than two times the bandwidth, a phenomenon known as aliasing of the analog signal will occur resulting in distortion of the analog waveforms. Even when the bandwidth of an ADC is properly matched to an application, aliasing can occur from an excessively wide bandwidth filter preceding the ADC that allows

broadband noise to pass into the ADC and/or signal harmonics which the ADC attempts to digitize.

It should be noted that, due to noise and bandwidth limitations, the effective number of bits (ENOB) is a specification often quoted by ADC manufacturers to represent a practical limit for the bit resolution of their devices. Ideally, an 8-b ADC would deliver all 8 b of resolution. Practical considerations, however, typically reduce an ideal 8-b ADC to an ENOB of 7.5 b.

For an integrated-circuit (IC) ADC, the analog input bandwidth is generally set by an on-board track-and-hold (T/H) amplifier preceding the quantizer circuitry. The MAX108 single-channel ADC from Maxim Integrated Products (Sunnyvale, CA), for example, is a 8-b device with sampling rate of 1.5 GSamples/s. It features an integrated T/H amplifier with 2.2-GHz full-power bandwidth and an on-chip +2.5-V precision bandgap voltage reference. The device achieves a near-ideal SNR of 46.8 dB and SFDR of –54 dBc. Designed for supply voltage of ±5 VDC, the ADC handles input signals over a range of ±250 mV. The accuracy of the component can be summarized in terms of its integral-nonlinearity (INL) and differential-nonlinearity (DNL) performance levels, which generally refer to the accuracy of an ADC's overall analog-to-dig-

High-speed-digitizer suppliers at a glance

COMPANY	PRODUCTS	WEBSITE
Acqiris	Digitizer cards	www.acqiris.com
Acquisition Logic	Digitizer boards	www.acquisitionlogic.com
Analog Devices	ADC ICs, boards	www.analog.com
Analogic	VXI instruments	www.analogic.com
DATEL	ADC ICs	www.datel.com
Echotek Corp.	Digitizer boards	www.echotek.com
Fairchild Semiconductor	ADC ICs	www.fairchild.com
Gage Applied	Digitizer boards	www.gage-applied.com
LeCroy Corp.	Digitizer instruments	www.lecroy.com
Linear Technology	ADC ICs	www.linear-tech.com
Maxim Integrated Products	ADC ICs	www.maxim-ic.com
Maxtek	Digitizer boards	www.maxtek.com
National Instruments	Digitizer instruments	www.ni.com
National Semiconductor	ADC ICs	www.national.com
Pentek	Digitizer boards	www.pentek.com
Philips Semiconductor	ADC ICs	www.semiconductors.philips.com
Rockwell Scientific	ADC ICs	www.rockwellscientific.com
Texas Instruments	ADC ICs	www.ti.com

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TYPICAL SPECIFICATIONS

Model	Freq (MHz)	Insertion Loss (dB)	Isolation (dB)	VSWR (:1)	Price $ea. Qty.10
•TCBT-2R5G	20-2500	0.35	44	1.1	8.95
•TCBT-6G NEW	50-6000	0.7	28	1.2	11.95

TCBT Actual Size .15"x.15" *LTCC*
•*Patent Pending*

					Qty.1-9
JEBT-4R2G	10-4200	0.6	40	1.1	39.95
JEBT-4R2GW	0.1-4200	0.6	40	1.1	59.95
PBTC-1G	10-1000	0.3	33	1.10	25.95
PBTC-3G	10-3000	0.3	30	1.13	35.95
PBTC-1GW	0.1-1000	0.3	33	1.10	35.95
PBTC-3GW	0.1-3000	0.3	30	1.13	46.95
ZFBT-4R2G	10-4200	0.6	40	1.13	59.95
ZFBT-6G	10-6000	0.6	40	1.13	79.95
ZFBT-4R2GW	0.1-4200	0.6	40	1.13	79.95
ZFBT-6GW	0.1-6000	0.6	40	1.13	89.95
ZFBT-4R2G-FT	10-4200	0.6	N/A	1.13	59.95
ZFBT-6G-FT	10-6000	0.6	N/A	1.13	79.95
ZFBT-4R2GW-FT	0.1-4200	0.6	N/A	1.13	79.95
ZFBT-6GW-FT	0.1-6000	0.6	N/A	1.13	89.95
ZNBT-60-1W	2.5-6000	0.6	45	1.10	82.95

NOTE: Isolation dB applies to DC to (RF) and DC to (RF+DC) ports

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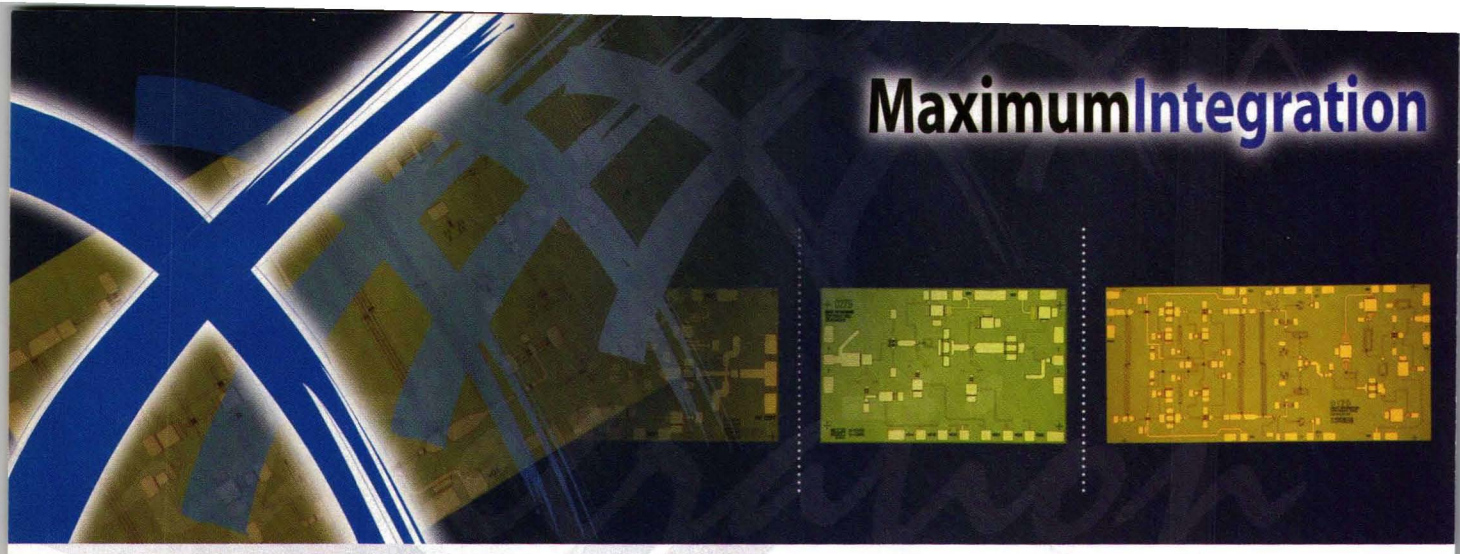

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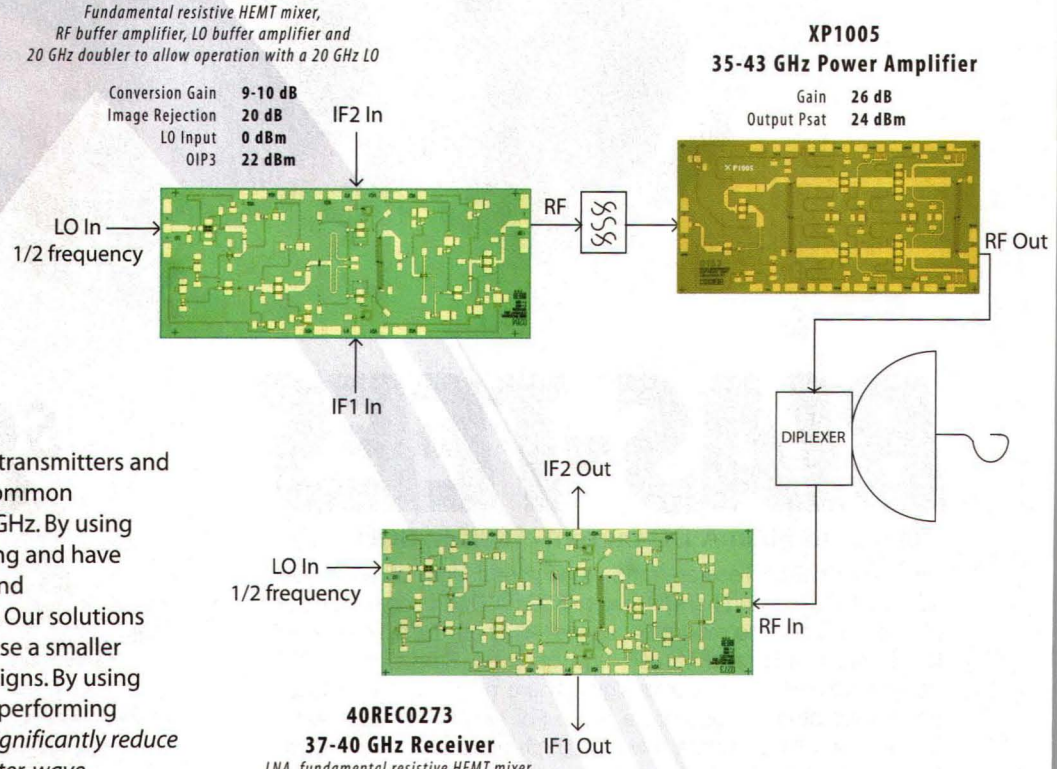
Conversion Gain	**9-10 dB**
Image Rejection	**20 dB**
LO Input	**0 dBm**
OIP3	**22 dBm**

XP1005
35-43 GHz Power Amplifier

Gain	**26 dB**
Output Psat	**24 dBm**



40REC0273
37-40 GHz Receiver
LNA, fundamental resistive HEMT mixer, LO buffer amplifier, 20 GHz doubler to allow operation with a 20 GHz LO

Noise Figure	**3.5-4.0 dB**
LO Input	**0 dBm**
Conversion Gain	**9-10 dB**
Image Rejection	**20 dB**
Input IP3	**5.0 dBm**

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ital transfer function and the accuracy of the step sizes in the ADC, respectively. Measured in terms of a range of an ADC's least-significant bit (LSB)—the smallest digital "slice" of the analog signal—the INL and DNL specifications are generally less than ±1 LSB to avoid ambiguities in the ADC's output data. In some cases, manufacturers will note "no missing codes" on their ADC data sheets to assure engineers that all digital codes are used. For the MAX108, the INL is ±0.25 LSB while the DNL is also ±0.25 LSB. The MAX108 is supplied in a 192-contact enhanced super ball-grid-array (ESBGA) housing. Lower-speed, pin-compatible versions of the device are also available, including the 1-GSamples/s model MAX104 and the 600-MSamples/s model MAX106.

Linear Technology (Milpitas, CA) also offers a family of high-speed ADCs in their LTC1740 series. The 12- and 14-b converters deliver SNR performance of better than 72 dB and SFDR performance of better than –85 dBc across wide input bandwidths. The model LTC1748, for example, is a 14-b ADC with 80 MSamples/s sampling rate. It achieves a 76.3-dB SNR at 5 MHz with a SFDR of –90 dBc at 5 MHz. As with the integrated devices from Maxim, the on-board T/H amplifiers dictate the maximum input bandwidths for these devices, and the LTC1748 offers among the widest bandwidth with a 240-MHz full-power T/H amplifier bandwidth.

Analog Devices (Norwood, MA) also supplies a wide range of high-speed ADCs, including the 14-b model AD6645 with 105 MSamples/s sampling rate. The monolithic converter includes a 200-MHz T/H amplifier and on-board voltage reference and maintains an SNR of 72 dB through the full bandwidth. The SFDR is 89 dBc for input signals to 70 MHz at sampling rates to 105 MSamples/s. The fourth-generation device is designed for multichannel, multimode receivers in which high multitone dynamic range is essential. The company's model AD9245 is a 14-b, 80-MSamples/s ADC that delivers 72 dB SNR and 85 dBc SFDR while consuming less than 500 mW power from a +3-VDC supply.

At lower-bit resolution, Analog Devices' model AD9410 is a 10-b, 210-MSamples/s ADC with 500-MHz analog bandwidth, on board voltage reference, integrated T/H amplifier, and demultiplexed outputs. Designed for +3.3 and +5 VDC supplies, it typically dissipates 2.1 W power and achieves an SNR of 54 dB with a 99-MHz input signal. The SFDR is 62 dBc for two-tone analog inputs at 80 and 81 MHz.

The company recently announced its AD92x9 family of quad ADCs: four converters fabricated on a single chip. Designed for space-constrained systems, such as medical imaging systems and multicarrier cellular base stations, and well suited for digitizing signals in multicarrier cellular base stations, the 12-b, 65-MSamples/s model AD9229 quad converter achieves a 70-dB SNR and 85-dBc SFDR with ±0.3 LSB DNL and ±0.6 LSB INL. It is designed for +3-VDC supplies and is supplied in a 48-pin LFCSP package. In addition, model AD9289 is an 8-b version in a ball-grid-array (BGA) package, with 47 dB SNR and 60 dBc SFDR. The ADCs feature serial low-voltage differential-signaling (LVDS) data outputs to simplify PCB layouts.

National Semiconductor (Santa Clara, CA) offers a 1-GHz ADC with their new ADC081000, an 8-b model with integral voltage reference and S/H amplifier. With DNL of ±0.25 LSB and power consumption of 1.4 W typical from a single +1.9-V supply, the converter employs an internal 1:2 demultiplexer to feed two LVDS buses and reduce differential output data on each bus to one-half the sampling rate. The ADC, which is ideal for direct downconversion in RF receivers, WLAN systems, and instrumentation and supplied in a 128-lead exposed-pad LQFP housing, features a power-down mode to reduce power consumption to typically 5 mW.

Texas Instruments (which acquired Burr-Brown) offers a 14-b ADC, model ADS5500, capable of sample rates to 125 MSamples/s. The single-channel device features an input bandwidth of 750 MHz, SNR of 70 dB, and SFDR of 82 dBc. The pipelined converter, which achieves INL of ±0.5 LSB, is designed for a +3-VDC supply and typically con-

sumes 750 mW power.

Some of the highest-speed ADCs currently on the market are GaAs monolithic devices from Rockwell Scientific Company, including the 6-b model RAD006, a 6-GSamples/s ADC. With an analog input bandwidth of 10 GHz, the device offers ENOB of 5.5 dB from DC to 3 GHz with SNR of better than 34 dB and SFDR or better than 49 dBc. With an INL of less than 1 LSB and DNL of less than 0.5 LSB, the high-speed ADC features a differential full scale range of 2 V peak to peak (pp). For those requiring slightly less speed but higher resolution, the company also offers the RAD008, with 8 b resolution and sampling rates to 3 GSamples/s (also a 10-GHz input bandwidth) and the 10-b model RAD0010, usable at sampling rates to 1 GSamples/s over a differential full-scale range of 1 V pp. The 8-b device offers SNR of better than 46 dB and SFDR of better than 67 dBc while the 10-b device has SNR of better than 55 dB. In addition to the companies mentioned, high-performance ADC ICs are also available from DATEL (Mansfield, MA), Fairchild Semiconductor (South Portland, ME), and Philips Semiconductors (Sunnyvale, CA).

When the digitizing function is needed along with supporting circuitry, such as memory and digital signal processing (DSP), it may make more sense to specify a board-level ADC rather than a chip. Maxtek, a subsidiary of Tektronix (Beaverton, OR), offers a variety of design and manufacturing services, including the use of thick-film ceramic and cofired multilayer-ceramic circuit boards, to create custom ADC-based signal-processing solutions.

In some cases, ADC cards and boards (also known as digitizers) resemble full-function measurement systems. The DBS908 digitizer from Analogic Corp. (Peabody, MA), for example, is a modular mezzanine board designed for use with the company's model DBS9905 C-size VXI carrier module. The digitizer features a sampling rate of 2 GSamples/s with 8-b resolution, 500-MHz analog input bandwidth, and 4-MSamples onboard memory. The single-channel DBS908 has an oscilloscope-type front end that can

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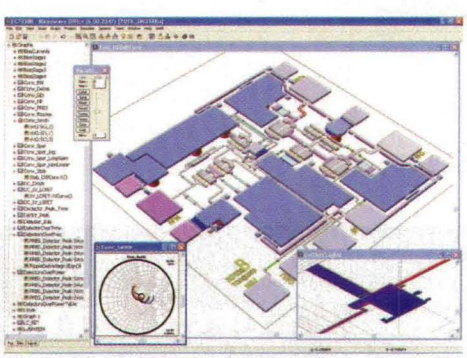

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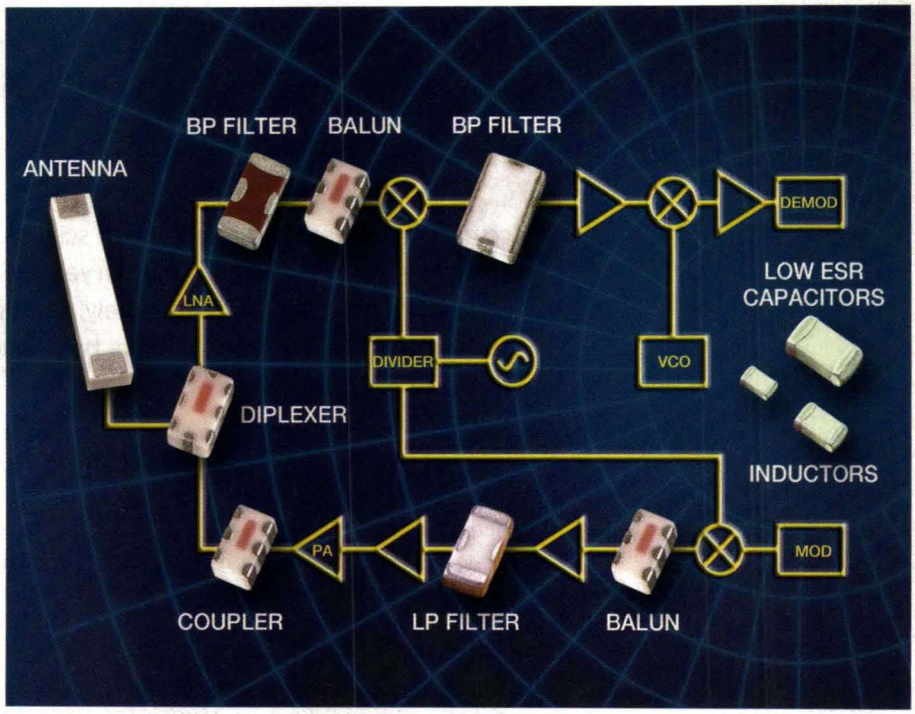
Microwave Office 2003

be programmed for AC or DC coupling and 50-Ω or 1 MΩ input impedance for a variety of measurement options. Signal acquisition can be initiated by triggering on the signal of interest (using an internal trigger), with an external trigger, or by VXI TTLTRG lines. By mounting several modules on the DBS9905 carrier, multiple signal-processing functions are possible from a single C-size VXI card. The DBS908 digitizer is well suited for automatic-test-equipment (ATE), mass spectrometry, and telecommunications testing applications.

Acqiris (Monroe, NY) is a Swiss company long associated with digitizer-on-board products. The company's model DC440 digitizer, for example, has been developed specifically for frequency-domain applications. It offers a 100-MHz DC-coupled standard input and a 300-MHz AC-coupled high-frequency input. Signals are captured on two channels at sample rates to 400 MSamples/s (per channel) and 12-b resolution. The board's ADC's achieve signal-to-noise ratios of better than 65 dB and spurious-free dynamic range (SFDR) of better than 80 dB. The digitizer is supplied on a PXI-compatible, 6U CompactPCI card for personal computers. The high-speed PCI bus can transfer data from the card to the computer at sustained rates to 100 Mb/s. The digitizer, which is supplied with device drivers for Windows 95/98/NT/2000/XP, VxWorks, and Linux, includes application code examples for C/C+, Visual Basic, and LabVIEW from National Instruments. The card consumes less than 25 W power.

LeCroy Corp. (Chestnut Ridge, NY) supplies the PXD series of PXI-based digitizers with bandwidths from 150 to 1000 MHz. Operating at sample rates to 2 GSamples/s, the cards are available with generous 256k acquisition memory (which can be increased as an option) and from 1 to 4 measurement channels. The PXD1022, for example, is a two-channel digitizer that fits into three PXI slots. It has a 1-GHz bandwidth, maximum single-shot sampling rate of 2 GSamples/s, and maximum repetitive sampling rate of 50 GSamples/s. The digitizer is ideal for commercial communications and military/aerospace test applications.

Additional board-level digitizers are listed in the **table**, including Echotek (Huntsville, AL), which offers the ECAD-1-081000 single-channel VMEbus ADC. This 8-b digitizer features a 1-GHz input bandwidth and 1 GSamples/s sampling rate with better than 42-dB SNR and 52 dBc SFDR. With trigger accuracy within two clock cycles, the digitizer board includes external analog, clock, and trigger input ports and as much as 16 MSamples of on-board memory storage, contained on a single-slot 6U VME module. MRF



Measurement Conference Tackles Differential Testing

The winter biannual meeting of this prestigious measurement society took on the challenge of making measurements on differential devices with single-ended gear.

JACK BROWNE
Publisher/Editor

Accuracy measurements require both science and skill. Knowledge of measurement equipment helps in testing, but often years of experience are needed to hone reliable measurement techniques. Fortunately, a dedicated group in the US, the Automatic RF Techniques Group (ARFTG), gathers twice each year to discuss measurement practices and techniques, and to share the wisdom of their members' experiences. The most recent meeting of ARFTG took place in the Hotel Boulderado in Boulder, CO, December 2-5 2003, with a focus on differential measurements.

The four-day ARFTG meeting consisted of a two-day short course on microwave measurements and instrumentation and the formal conference on differential measurements. The short course, as might be expected for a measurement meeting held in Boulder, was dominated by speakers from the nearby National Institute for Standards and Technology (NIST). NIST speakers included John Juroshek (on RF connectors and transmission lines), Tom Crowley (on microwave power measurements), Dylan Williams (on achieving accurate on-wafer measurements), Tracy Clement (on practical oscilloscope measurements), Paul Hale (with an introduction to optoelectronic measurements), and Jim Randa (on thermal noise measurements).

The NIST reunion was joined by long-time ARFTG member Doug Rytting of Agilent Technologies (Santa Rosa, CA) who spoke on vector-network-analyzer (VNA) error models and calibration methods, Nick Ridler of the National Physics Laboratory (Malverne, England) who discussed uncertainty analysis, Ed Godshalk of Maxim Integrated Products (Beaverton, OR) who covered phase-noise measurements, Jan Verspecht of Jan Verspecht bvba who reviewed on large-signal network analysis, and Howard Reader of the University of Stellenbosch who detailed how to pay attention to electromagnetic compatibility (EMC) in the measurement laboratory.

The differential measurements conference, held during December 4-5, 2003, was chaired by NIST's Dylan Williams who was assisted by NIST's Kate Remley. The conference included sessions on differential measurements, temporal measurements for 40-Gb/s optical-communications systems, nonlinear measurements, circuit and device modeling, and VNA calibrations. In the differential measurement sessions, for example, James Broomall and associates from W.L. Gore & Associates (Elkton, MD) offered practical information on a coaxial-to-differential adaptor capable of making the transition from single-ended 3.5-mm coaxial devices to differential devices. In the same session, Yves Rolain and colleagues from Vrije Universiteit Brussel (Pleinlaan, Brussels, Belgium) explained how to measure the nonlinear characteristics of differential RF amplifiers using a single-ended source. Also, Kipp Schoen of Picosecond Pulse Labs (Boulder, CO) reported on a sub-10-ps differential pulse source for time-domain-reflectometer (TDR) and time-domain-transmission (TDT) measurements.

The next (63rd) meeting of ARFTG, which is dedicated to "on-wafer characterization," is scheduled for June 11, 2004 in Fort Worth, TX in conjunction with the IEEE Microwave Theory & Techniques Symposium (MTT-S). For more information on ARFTG itself or on attending the 63rd meeting, visit the website at www.arftg.org. MRF

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3WAY	0.50-4.20
4WAY	0.47-8.40
5WAY	0.50-1.98
6WAY	0.80-5.00
7WAY	0.85-1.99
8WAY	0.50-8.40
9WAY	0.80-4.80
10WAY	0.75-2.40
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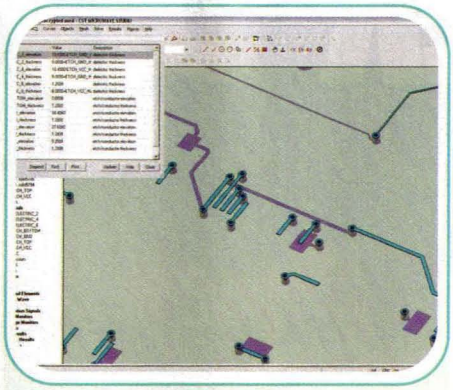
Software Tools Feature Enhanced EDA Workflow

DESIGNERS OF HIGH-FREQUENCY packages and PCBs who rely on the Allegro suite of software tools (formerly known as the Allegro PCB/APD tools) from Cadence Design Systems (San Jose, CA) can now benefit from a direct interface to the Microwave Studio suite of simulation tools from Computer Simulation Technology (CST). The link allows signal-integrity studies to be performed on package and PCB designs, taking into account all 3D field effects. The new capability is available in Version 5.0.1 and later versions of the Microwave Studio software suite, a related group of software tools that supports the time-domain EM analysis of high-speed/high-frequency circuits.

Computer Simulation Technology (CST), 10 Laurel Ave., Suite 300, Wellesley Hills, MA 02481; (781) 416-2782, FAX: (781) 416-4001, e-mail: info@cst-america.com, Internet: www.cst.com.



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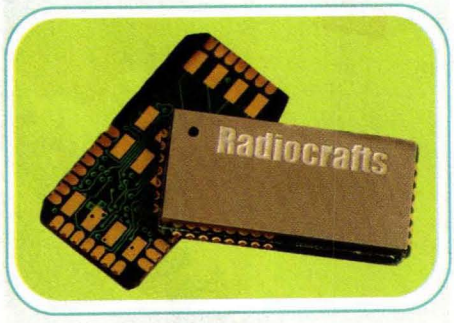
High-Speed Converter Digitizes 200 MHz

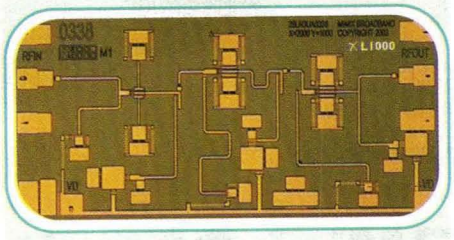
WELL SUITED FOR multichannel, multimode cellular receivers and base stations, the AD6645 high-speed ADC provides 14-b precision at sampling rates to 105 MSamples/s. It is capable of accurately converting differential-analog IF signals as high as MHz to 14-b digital code in the form of 3.3-V CMOS-compatible digital outputs. Fabricated with a high-speed complementary bipolar process, the ADC achieves a signal-to-noise ratio (SNR) of 75 dB at a sampling rate of 105 MSamples/s and input frequency of 15 MHz and an SNR of 72 dB at a sampling rate of 105 MSamples/s and input frequency of 200 MHz. The converter features sampling jitter of less than 0.1 ps, power dissipation of only 1.5 W. It is rated for operating temperatures from –40 to +85°C.

Analog Devices, Inc., One Technology Way, P.O. Box 9106, Norwood, MA 02062-9106; (617) 329-4700, Internet: www.analog.com.

ANALOG DEVICES' AD6645 HIGH-SPEED ADC



RADIOCRAFTS AS' SERIES OF RF MODULES



MIMIX BROADBAND'S MODEL XL1000 LOW-NOISE AMPLIFIER

RF Modules Have Embedded Protocol

COMPACT BUT VERSATILE, the RC1040, RC1080, and RC1090 series of RF modules blend complete RF data transceivers with an embedded protocol for ease of installation in short-range wireless products. Models RC1040 and RC1080 are designed for operation in the license-free bands in Europe at 433 and 868 MHz, respectively, while the model RC1090 is designed for use in the United States and Canada under FCC regulation in the 902-to-928-MHz Industrial-Scientific-Medical (ISM) band. The modules achieve sensitivity of –100 dBm for data rates to 19.2 kb/s and produce transmitter output-power levels to +10 dBm. The fully shielded and surface-mountable modules measure only 12.7 × 25.4 × 3.5 mm.

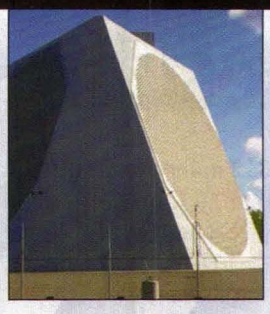
Radiocrafts AS, Gunnar Schjelderups vei 11, NO-0485 Oslo, Norway; (47) 970 86 676, e-mail: sales@radiocrafts.com, Internet: www.radiocrafts.com.

Self-Biasing LNA Spans 20 To 40 MHz

BASED ON A GaAs MMIC three-stage architecture, the model XL1000 low-noise amplifier covers the full octave from 20 to 40 GHz with a fullband noise figure of 2.5 dB and small-signal gain of 20 dB. The rugged high-electron-mobility-transistor (HEMT) amplifier IC, which can be powered by a single +3- or +5-VDC bias supply with no need for negative voltage or active bias circuitry, achieves output power at 1-dB compression of +8 dBm. The amplifier is well suited for wireless-communications applications including millimeter-wave point-to-point radios, local-multipoint-distribution-services (LMDS) systems, satellite-communications systems, and millimeter-wave radar systems. The chip, which undergoes 100-percent on-wafer testing, features surface passivation for environmental protection and gold metallization for superior performance at high frequencies.

Mimix Broadband, Inc, 10795 Rockey Rd., Houston, TX 77099; (281) 988-4600, FAX: (281) 988-4615, Internet: www.mimixbroadband.com.

M/ACOM



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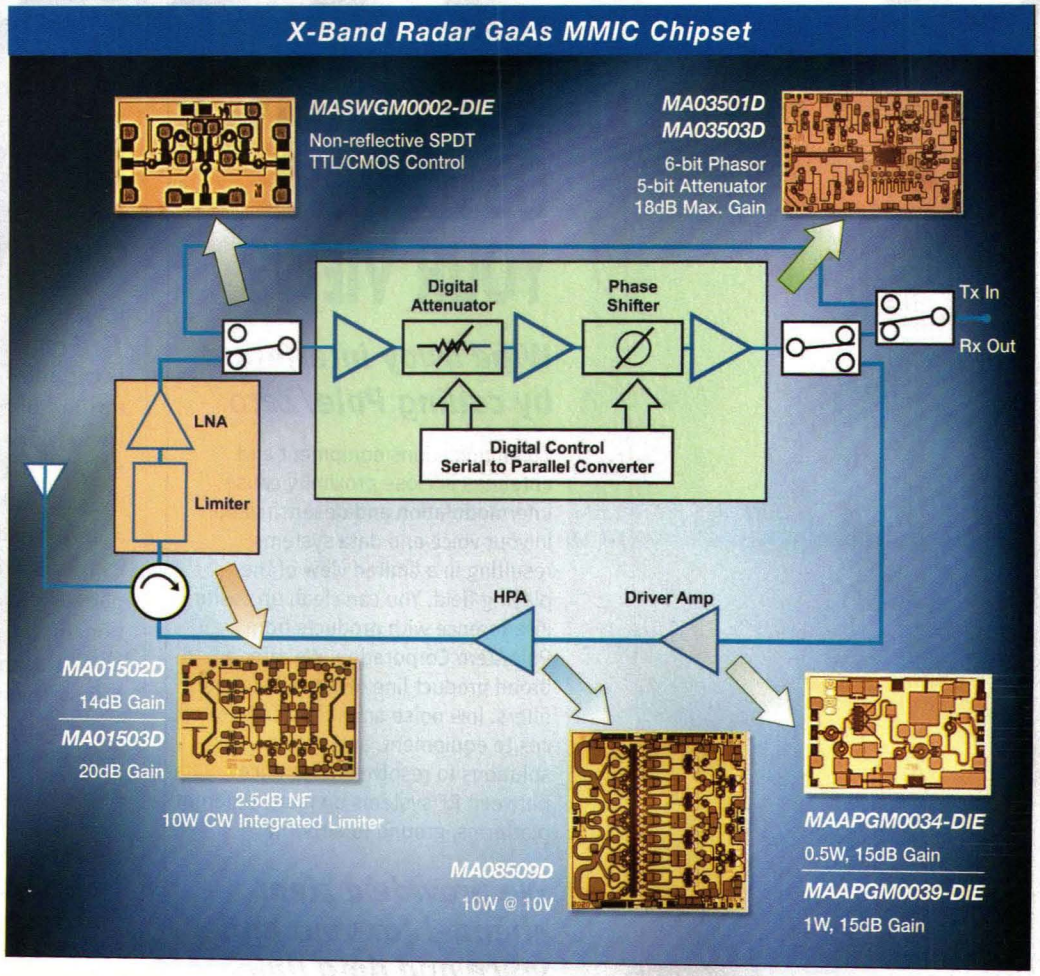

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QUALCOMM Announces Q1 Results

QUALCOMM, INC. announced its first-quarter fiscal 2004 results ended December 28, 2003. Revenues were $1.2 billion in the first fiscal quarter, up 37 percent sequentially and 13 percent year-over-year. First fiscal quarter net income was $352 million and earnings per share were $0.43, up 21 percent and 23 percent sequentially and up 46 percent and 43 percent year-over-year, respectively.

Revenues excluding the QUALCOMM Strategic Initiatives (QSI) segment were $1.2 billion in the first fiscal quarter, up 39 percent sequentially and 13 percent year-over-year. First fiscal quarter net income excluding the QSI segment was $419 million and earnings per share were $0.51, up 77 percent and 76 percent, respectively, compared to the prior quarter, and both up 21 percent year-over-year.

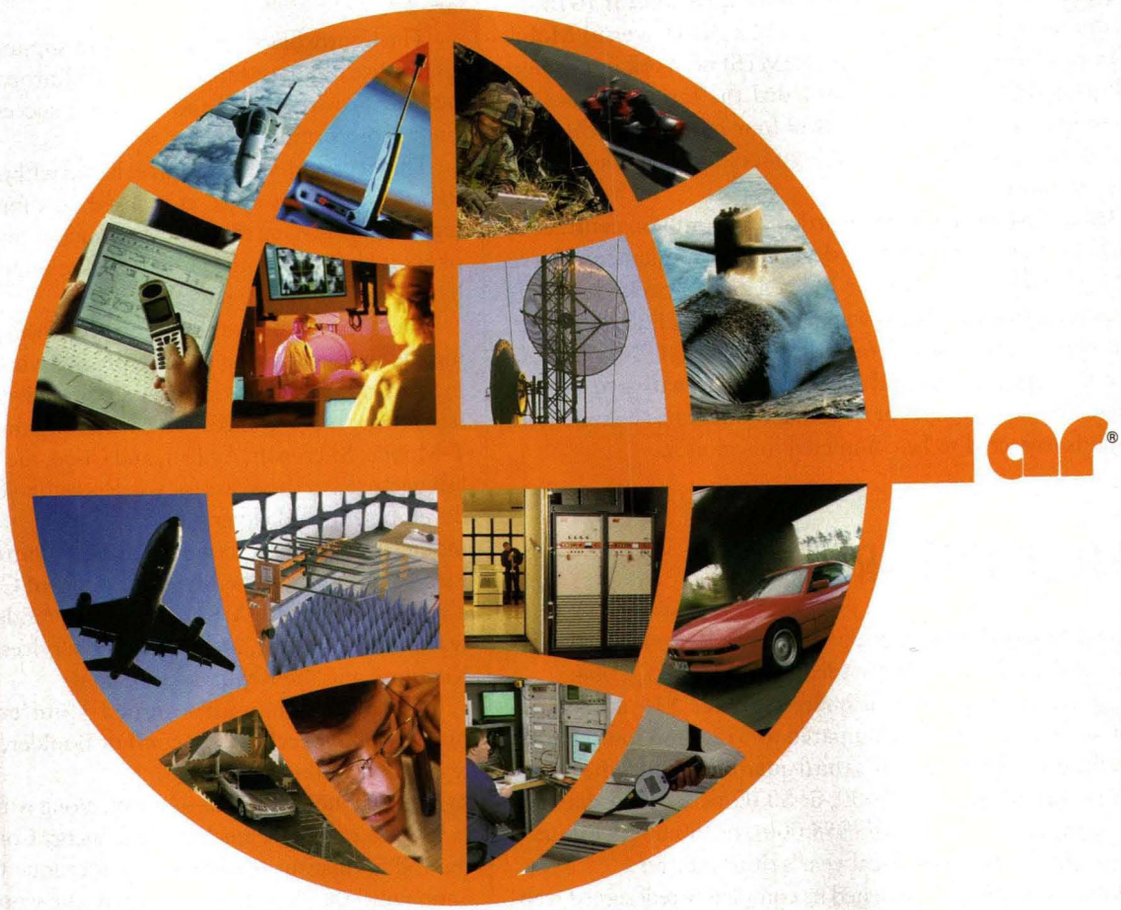
"Our results for the first quarter of fiscal 2004 have exceeded our expectations by every measure," says Dr. Irwin Mark Jacobs, chairman and CEO of QUALCOMM. "We now anticipate the full 2004 fiscal year to be stronger than we had earlier anticipated due to increased worldwide demand for CDMA phones and devices with increased functionality, such as color screens, cameras, and multimedia capabilities. We also believe that CDMA channel inventories approached near record low levels in the fourth fiscal quarter, fueling increased demand for our MSMs."

Research-and-development (R&D) expenses were $150 million in the first fiscal quarter, up 33 percent year-over-year, largely attributable to increased R&D investment in QUALCOMM CDMA Technologies (QCT) to continue to drive broadly segmented chipset platforms in response to market demand including WCDMA (UMTS), CDMA2000 1X, and 1xEV-DO.

Selling, general, and administrative (SG&A) expenses were $137 million in the first fiscal quarter, down 8 percent year-over-year, largely attributable to a decrease in SG&A expenses of the Vesper Operating Companies in Brazil, which QUALCOMM sold during the fiscal first quarter.

Further information can be found at www.qualcomm.com. MRF

AR's Global Competitive Edge.



Times have changed; it's no longer business as usual. Companies are looking for that global competitive edge. To compete in today's environment, you have to have more than a myopic view of what you deliver.

At AR Worldwide, Quality must = Value and AR Value is multi-dimensional. With advanced design like our 3000W1000 high-power amplifier to low-power amplifiers with infinite mismatch tolerance, AR Worldwide is there. Innovative products like our Radiant Arrow "bent element" family of antennas are an industry breakthrough. Our "Vision Concept" approach to RF modules offers speed to market and tremendous value. And, we meet or exceed the most stringent standards and specifications the world over.

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CONTRACTS

Motorola's Global Telecom Solutions Sector (GTSS)—Announced an agreement to build a 3G Universal Mobile Telecommunications System (UMTS) network in northern Portugal. Motorola was awarded the three-year deal by Optimus following a successful four-month network trial. Optimus plans to deliver 3G services to northern Portugal by Q2 2004.

Harris Corp.—Has been awarded a $24 million contract by MTN Nigeria to supply, design, and implement Harris MegaStar® 155 radios for MTN's high-capacity backhaul GSM network. MTN is one of the largest cellular networks in Africa. The new Harris radios provide MTN with higher network capacity and a lower cost alternative to existing satellite links. Shipment of the equipment has commenced, and is expected to be completed this month.

FRESH STARTS

Ansoft Corp.—Announced financial results for its third quarter of fiscal 2004, which ended on January 31, 2004.

Revenue for the third quarter totaled $14 million, an increase of 13 percent compared to $12.4 million reported in the previous fiscal year's third quarter. Net income for the third quarter was $941,000, or $0.07 per diluted share, as compared to a net loss of ($58,000), or ($0.00) per diluted share in the previous fiscal year's third quarter.

Noble Publishing—Launched its completely redesigned website, which features books, instructional CD-ROMs, and software for the practicing RF and microwave engineer. The site includes a new shopping cart, enhanced navigational and search capabilities, and discounts on selected book and CD-ROM packages. The site is located at www.noblepub.com.

SSB Technologies, Inc.—Announced that effective January 1, 2004 they have reincorporated under their Stealth Microwave trademark name. Stealth Microwave, Inc. is an ISO-9001:2000-certified company that provides ultra-linear single and multi-carrier RF and microwave amplifiers to wireless and military markets. Their new website can be found at www.stealthmicrowave.com.

Amplifier Research (AR)—Has changed its name to AR Worldwide.

For more information, contact: AR Worldwide, 160 School House Rd., Souderton, PA 18964; (215) 723-8181, Internet: www.ar-worldwide.com.

ARKalmus—Has changed its name to AR Northwest. The renamed company will also have a new general manager. Chris Heavens has joined parent company AR Worldwide as a vice president who will oversee operations at AR Northwest.

For more information, contact AR Northwest, 11807 North Creek Parkway South, Suite 109, Bothell, WA 98011; (425) 485-9000, Internet: www.ar-northwest.com.

Wireless Innovation Centre—Officially launched on February 19.

The Centre has been created with support from Scottish Enterprise Renfrewshire, Strathclyde European Partnership, and the private sector, and follows a successful six-month pilot project run in 2003.

This Scotland-wide resource, based within the Hillington Park Innovation Centre, is focused on supporting, nurturing, and encouraging the conception and sustained development of Scottish SMEs in the global market opportunity offered by the benefits of wireless technology.

The Wireless Innovation Centre is a focal point for growing the first generation of enterprising wireless companies in Scotland. With links into key stockholders in the Scottish technology marketplace, and partner organizations such as T-Mobile, Microsoft, Agilent, and Cisco, the Centre is establishing itself as a showcase for wireless technology, application development, and expertise.

Modelithics, Inc.—Announced the appointment of Melcom Electronics Ltd. (www.Melcom.co.uk) of Chertsey, UK as a reseller for its EDA RF/microwave model library software, measurement, and modeling services in the United Kingdom and the Republic of Ireland.

Modelithics has also signed Conifer Technology (www.ConiferTechnology.com) of Boulder, CO as its representative in the western US.

Agilent Technologies, Inc.—Agilent, along with the CEA-Leti (a laboratory of the French Atomic Energy Commission), and the INPG (Institut National Polytechnique Grenoble), has announced the joint development of a new optical and high-speed communications test and design facility in Grenoble, France. The PLATEST (platform and test) facility enables startup companies and R&D organizations to develop communications components and subsystems. These companies benefit by leasing the facility rather than purchasing new test and measurement instruments.

The facility is a development within the Minatec Innovation Centre, the largest European research center for micro- and nanotechnologies, that is now under construction in Grenoble.

Credence Systems Corp.—Announced that Micrel, Inc. has purchased multiple ASL 3000TM and ASL 1000TM mixed-signal test systems. Micrel will use the systems for high-volume production test of its high-performance linear and mixed-signal integrated circuits (ICs). Micrel chose the ASL systems based on their ability to achieve a lower overall cost-of-test than competitive systems.

Tanner EDA—Has opened a wholly owned subsidiary in Hsinchu, Taiwan to accommodate increasing customer demand in Asia for cost-effective, Windows-based EDA tools. This is Tanner's second office in the region, following the establishment of Tanner Research Japan in April 2003. MRF

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Mini-Circuits is pleased to announce the opening of additional MITRON, INC. sales offices serving greater China. As specialists in every aspect of customer service, you can depend on Mitron, Inc. for Mini-Circuits current and new product information, answers to your technical and application questions, price quotes, order communications, shipment confirmations, literature and publications, and much more! No matter what your RF, IF, & Microwave needs may be, the sales professionals at Mitron, Inc. will provide you with the highest level of customer service. Now, one simple email or call to any one of 7 conveniently located offices connects you directly to Mini-Circuits large variety of products and services. Just one more way we're working hard to help you become even more successful!

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See our 244 page RF/IF Designer's Guide in EEM (Electronic Engineers Master)


BROWN

Brown Is Named To Sales Position At Spectrum Control

MARGO BROWN has joined Spectrum Control as director of channel sales for the company's Power Management System Group. Prior to joining Spectrum Control, Brown worked as a consultant with Cisco Systems and was employed in sales at Sonicblue.

Eagleware Corp.—MARK SHUFFIELD to regional sales director; formerly worked at AWR and Agilent's EEsof group. Also, SCOTT MAYNARD to regional sales director; formerly employed with AWR.

Green Hills Software, Inc.—DANIEL P. BURNHAM to the board of directors; formerly chairman and CEO of Raytheon Co.

ABI—ALAN VARGHESE to senior director of wireless research; formerly director of engineering for transmission network systems at Scientific-Atlanta. Also, RAY JODOIN to director of wireless infrastructure research; formerly director of wireless research for In-Stat.

The Aerospace Industries Association—GINETTE C. COLOT to corporate secretary, treasurer, and CFO; formerly vice president of administration at Goodwill Industries International.

Harris Corp.—LEON V. SHIVAMBER to corporate vice president for supply chain management and operations; formerly president of Vertisync, Inc.

Fujitsu Compound Semiconductor, Inc. (FCSI)—JANE LI to vice president for lightwave products; formerly employed in executive-level positions at NeoPhotonics, Novalux, Inc., and Corning.

Leap Wireless International, Inc.—S. DOUGLAS HUTCHESON to executive vice president and CFO; formerly senior vice president and CFO. Also, GLENN UMETSU to executive vice president and COO; formerly senior vice president for engineering operations.

Texas Instruments, Inc.—JEAN-FRANÇOIS FAU to president of TI Europe; formerly had overall management responsibility for the European wireless-communications business unit.

IceFyre Semiconductor, Inc.—GREG STANLEY to vice president of operations; formerly vice president of operations at SiberCore Technologies, Inc.

Cookson Electronics—PETER A. PALMER to senior vice president of marketing and new business development; formerly vice president of marketing for Cookson Electronics Assembly Materials.

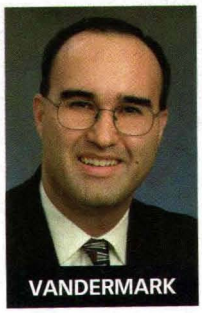
Aruba Wireless—DON LEBEAU to president and CEO; formerly senior vice president of worldwide sales and operations for Cisco Systems.

California Eastern Laboratories—PAUL MINTON to executive vice president; formerly vice president for corporate development.

Megaphase—BOB FISHER to chief engineer; formerly employed at ITT Cannon, FCI, and AMP.

KDF—DR. AMMAR DERRAA to director of technology; formerly led a nanotechnology project at NanoOpto Corp.


DERRAA


VANDERMARK

Brush Ceramic Products, Inc.—LEE VANDERMARK to sales and marketing manager; formerly employed in sales and marketing positions.

Wireless Valley Communications, Inc.—JIM WELCH to president and CEO; formerly president and CEO of Metrowerks.

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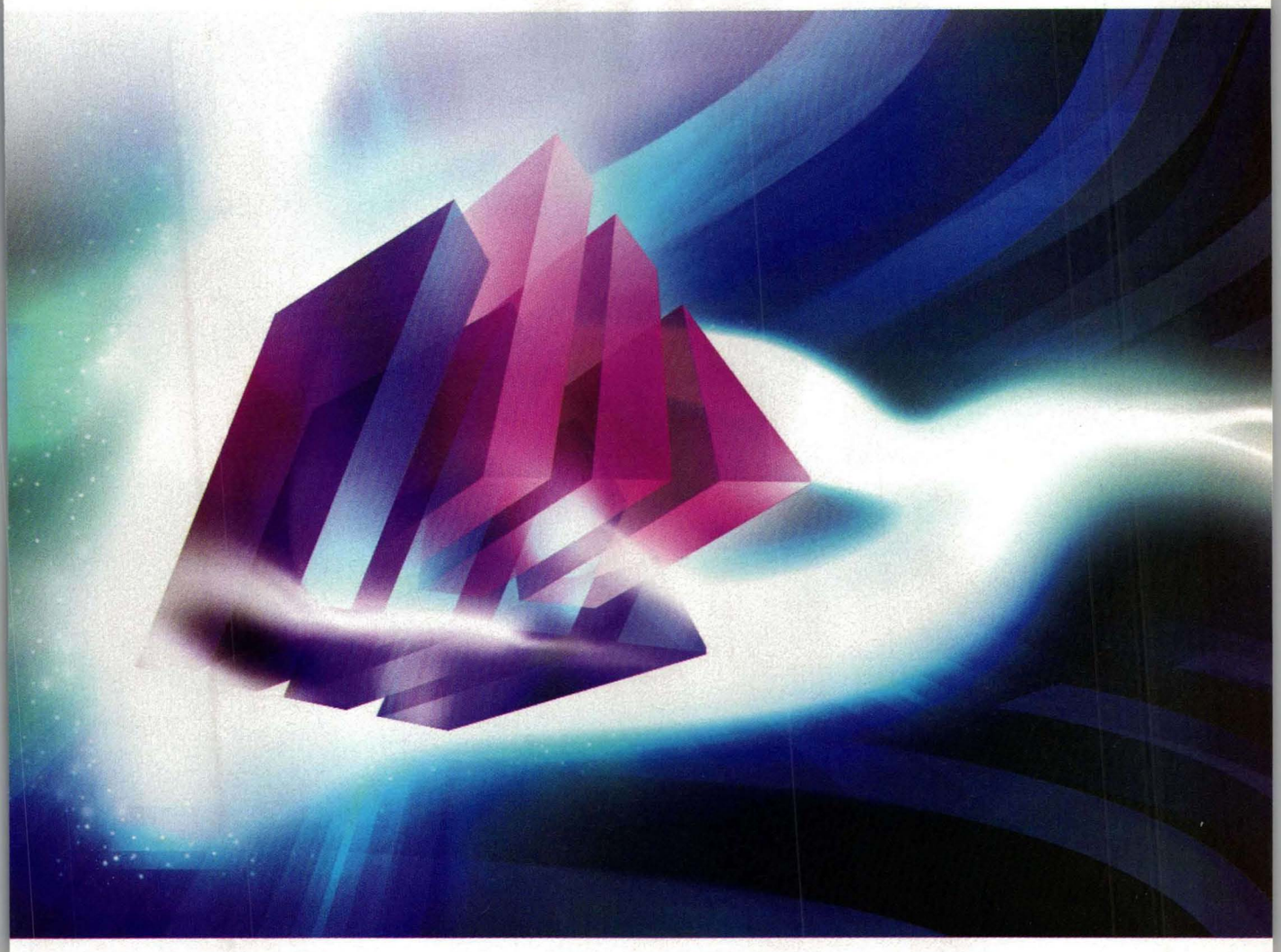

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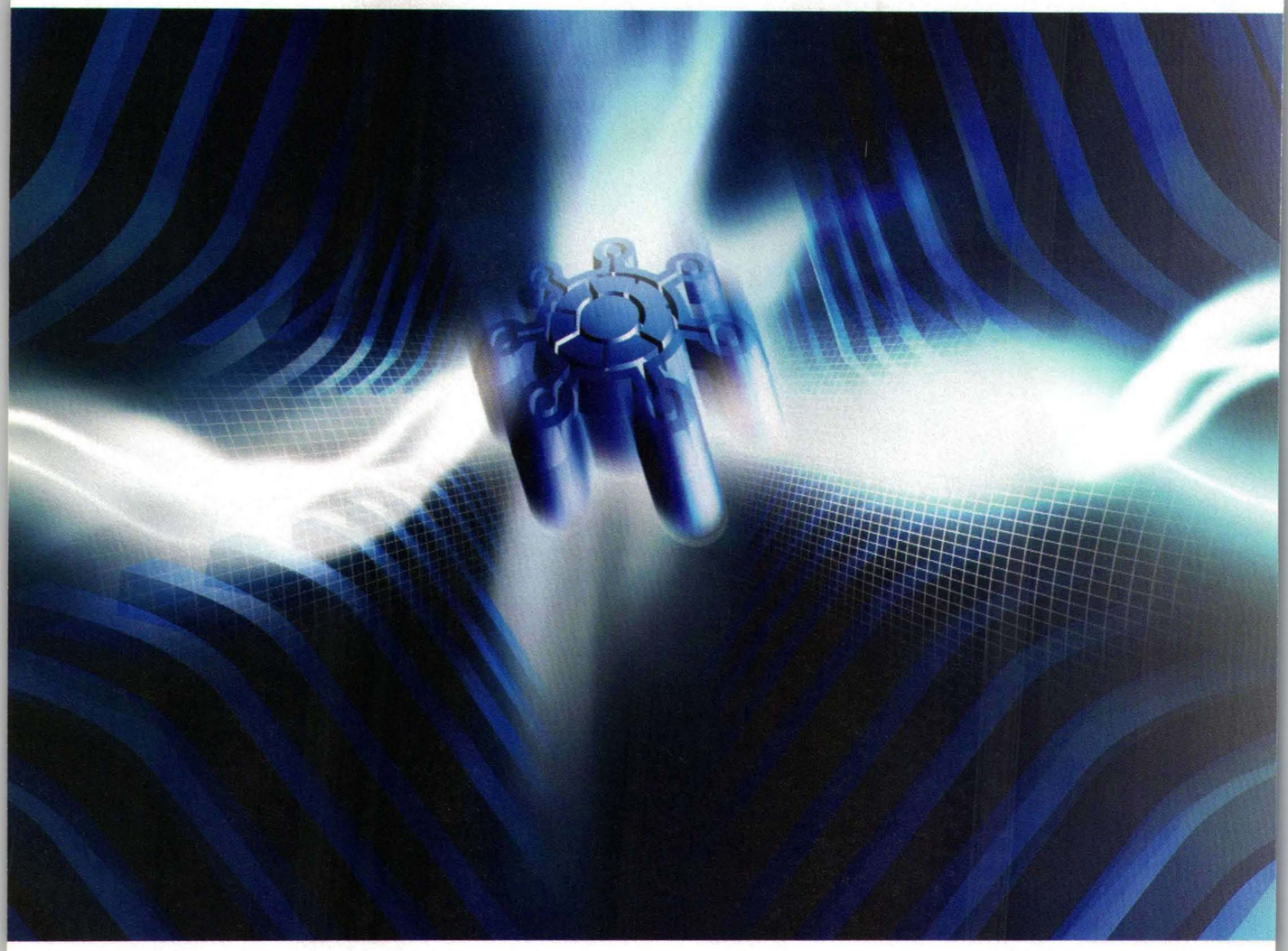

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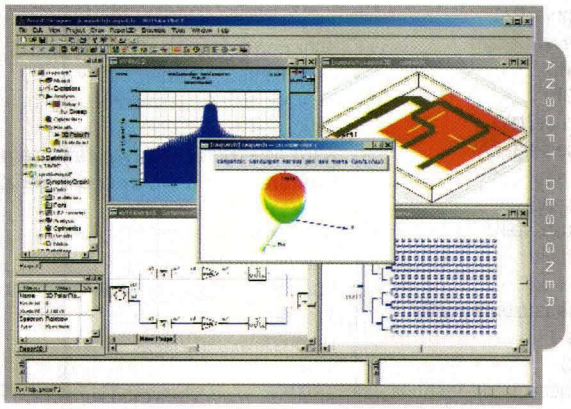

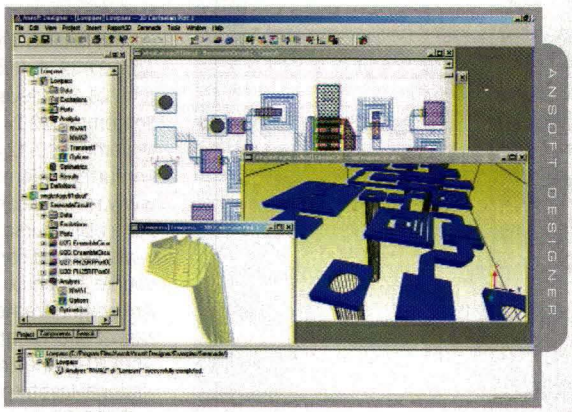

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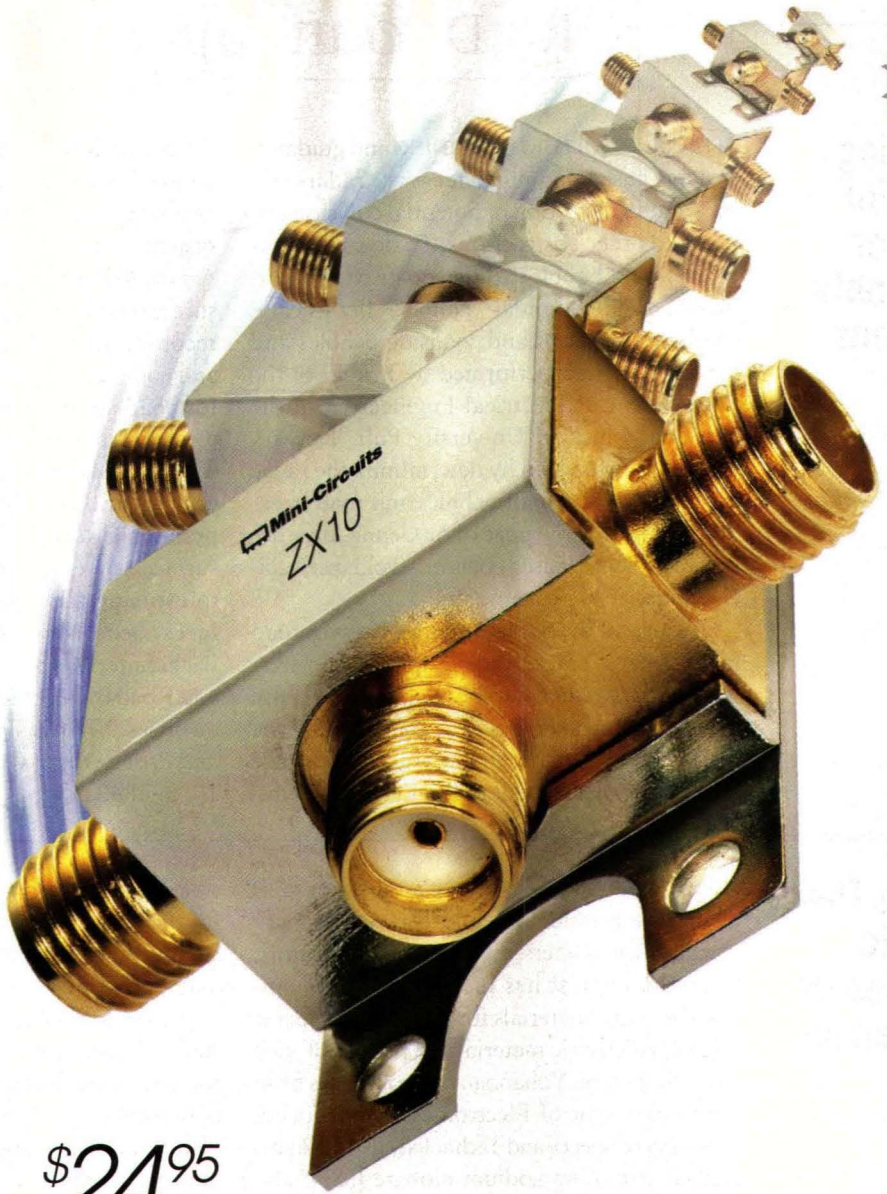

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Developing Models For Millimeter-Wave Pebble Reflections

VEHICULAR COLLISION-WARNING and guidance systems based on millimeter-wave radars rely on accurate detection of potential collision targets and other obstructions. To do so, however, they must be able to differentiate what is normally considered radar "clutter," in this case, road-surface faults and roadside pebbles and debris. Research performed by Eric Li of the Department of Electrical Engineering of the National Chi Nan University Puli, Taiwan, ROC was motivated by determining the radar cross section (RCS) of pebbles and road irregularities for the purpose of modeling in future millimeter-wave-based collision-avoidance/vehicle-guidance systems.

Because of the backscattering effects of road faults, such as cracks and potholes, millimeter-wave radar systems can be triggered into false alarms. As a result, it is desirable to study the RCS of such road defects in order to improve the accuracy of vehicular radar-tracking systems.

Research was based on the physical optics approximation of the backscatter response of roadside debris. A 94-GHz polarimetric radar system from the University of Michigan (Ann Arbor, MI) was used for the backscatter measurements, with a variety of incident angles recorded for different-sized roadside cracks and potholes. Statisitical models were developed for road-surface roughness and a variety of theoretical models were developed and compared with the polarimetric measurements. The integral equation method (IEM) was applied to predict the surface scattering from the pebble surface while a hybrid model was employed to estimate the scattering from smooth serfaces. See "Physical Optics Models for the Backscatter Response of Road-Surface Faults and Roadside Pebbles at Millimeter-Wave Frequencies," *IEEE Transactions on Antennas and Propagation*, October 2003, Vol. 51, No. 10, p. 2862.

Studying The Dielectric Properties Of SSN Ceramics

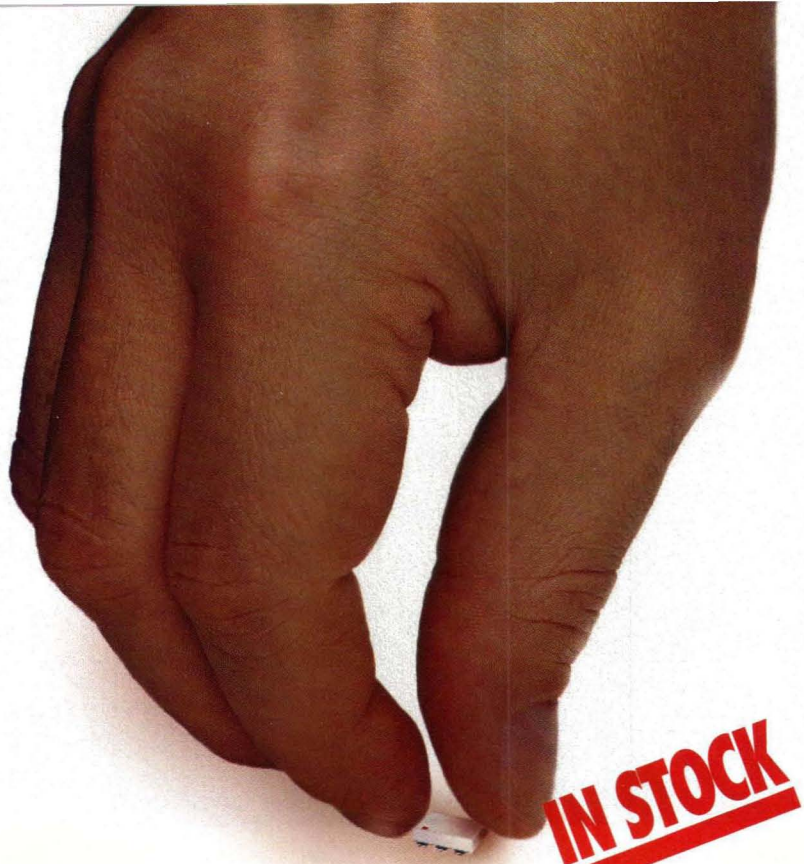
CERAMIC MATERIALS HAVE LONG BEEN used as dielectric substrates for high-frequency resonant circuits, such as filters, diplexers, and resonators. Recently, interest has increased in niobium-based ceramic materials for high-quality-factor (high-Q) dielectric materials. In pursuit of such research, Jaimon Yohannan and colleagues from the Department of Electronics at the Cochin University of Science and Technology (Kochi, India) studied strontium sodium niobate (SSN) and numerous other high-Q ceramic materials for their suitability as microwave substrates.

The researchers employed the cavity perturbation technique for their study. In this approach, a closed section of waveguide constitutes a waveguide cavity resonator. Electromagnetic (EM) energy is coupled to the cavity through coupling irises at the ends of the cavity. A nonradiating slot is provided at the broad wall of the cavity for the introduction of a dielectric sample. The cavity resonates at different frequencies, depending upon its physical dimensions.

A model 8714ET vector network analyzer from Agilent Technologies (Santa Rosa, CA) was used for measuring the resonant-frequency spectrum of the samples under test. Several resonant peaks were found in the range from 2 to 3 GHz, including approximately 2.45 and 2.7 GHz, while the conductivity was found to vary for different samples of SNN material. In general, the dielectric constant was found to decrease with increasing frequency. See "Dielectric Properties of Strontium Sodium Niobate Ceramics Using the Microwave Cavity Perturbation Technique," *Microwave and Optical Technology Letters*, October 20, 2003, Vol. 39, No. 2, p. 112.

CMOS Fractional-N Synthesizer Reaches 1.8 GHz

FRACTIONAL-N FREQUENCY synthesizers are highly regarded for their overall RF performance. In support of wireless applications, Chun-Huat Heng, a student at the University of Illinois (Champaign, IL) and B.-S. Song of the Department of Electrical and Computer Engineering at the University of California at San Diego (La Jolla, CA) developed a 1.8-GHz fractional-N frequency synthesizer based on a simple silicon CMOS process. The architecture includes a multiphase voltage-controlled oscillator (VCO) and multiphase frequency divider. Operating from a +3.3-VDC supply, the synthesizer achieved a tuning range of 1.675 to 1.795 GHz with resolution as good as 10 Hz and single-sideband phase noise of –118 dBc/Hz offset 1 MHz from a 1.715-GHz carrier. Using a 20-MHz reference frequency, the source maintained spurious tones at –70 dBc while consuming only 52 mW of power. See "A 1.8-GHz CMOS Fractional-N Frequency Synthesizer With Randomized Multiphase VCO," *IEEE Journal of Solid-State Circuits*, June 2003, Vol. 38, No. 6, p. 848.



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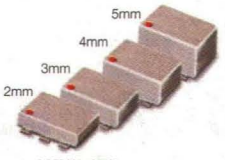

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ADE-1	+7	0.5-500	5.0	55	15	4	1.99▲
ADE-1ASK	+7	2-600	5.3	50	16	3	3.95
ADE-2	+7	5-1000	6.67	47	20	3	1.99▲
ADE-2ASK	+7	1-1000	5.4	45	12	3	4.25
ADE-6	+7	0.05-250	4.6	40	10	5	4.95
ADEX-10	+7	10-1000	6.8	60	16	3	2.95
ADE-12	+7	50-1000	7.0	35	17	2	2.95
ADE-4	+7	200-1000	6.8	53	15	3	4.25
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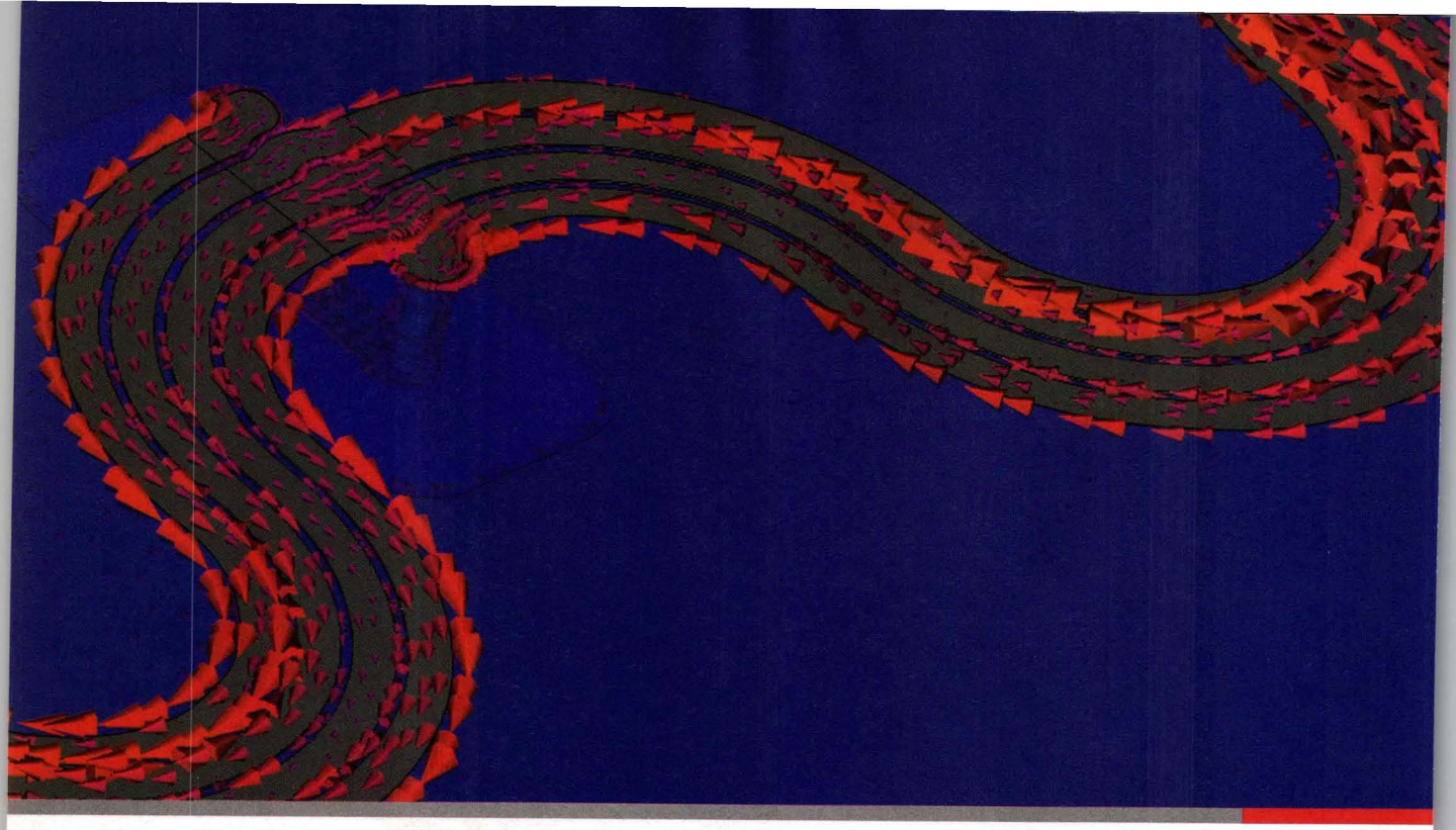

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Steering Through RKE Requirements

Understanding regulatory-agency requirements and the limitations of modern device technologies can simplify the task of designing short-range remote-keyless-entry (RKE) systems.

Remote-keyless-entry (RKE) capability has captivated automotive buyers, with an RKE installation rate of more than 80 percent for new vehicles in North America and more than 70 percent in Europe. Most RKE systems employ one-way (simplex) communications (from the key to the door lock, for example), although second- and third-generation RKE systems may incorporate duplex operation with communication back to the key. Designing an effective RKE system involves understanding emissions limits set by the Federal Communications Commission (FCC) as well as technical capabilities of integrated circuits (ICs) and supporting circuitry.

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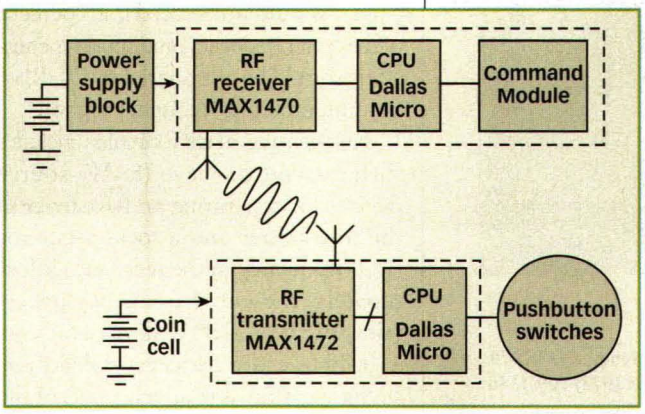
An RKE system consists of an RF transmitter in the key fob (or key) that sends a short burst of digital data to a receiver in the vehicle, where it is decoded and made to open or close the vehicle doors or trunk via receiver-controlled actuators. The wireless link is simply a carrier frequency, currently 315 MHz in the United States and Japan, and 433.92 MHz of the Industrial-Scientific-Medical (ISM) band in Europe. Japanese RKE systems employ frequency-shift-keying (FSK) modulation, but in most other parts of the world amplitude-shift-keying (ASK) modulation is used, in which the carrier is amplitude modulated between two levels. To save power, the lower level is usually near zero, producing complete on-off-keying (OOK) modulation.



1. An RKE system consists of a key fob circuit (lower diagram) transmitting to a receiver in the vehicle (upper diagram).

Typical RKE systems **(Fig. 1)** include a microcontroller in the key or key fob. A car is typically unlocked by pressing a pushbutton on the key that wakes up the microcontroller and sends a 64- or 128-b data stream to the key's RF transmitter, where it modulates the carrier and is radiated via a simple printed-circuit loop antenna. (Although inefficient, a loop antenna fabricated as part of the printed-circuit board (PCB) is inexpensive and widely used.) In the vehicle, an RF receiver captures that data and directs it to another microcontroller, which decodes the data and sends an appropriate message to start the engine or open the door. Key fobs with multiple buttons provide choices of opening the driver's door, or all doors, or the trunk, etc.

The RKE digital data stream, transmitted between 2.4 and 20 kb/s, usually consists of a data preamble, a command code, some check bits, and a "rolling code" that ensures vehicle secu-

rity by altering itself with each use. (Otherwise, a transmitted signal might accidentally unlock another vehicle, or fall into the hands of a car thief who could use it to gain entry later on.)

Several major objectives govern the design of these RKE systems. Like all mass-produced automotive components, they must offer low cost and high reliability. They should minimize power drain in both transmitter and receiver, because replacing batteries in a key fob is a nuisance and recharging the car battery is a major nuisance. With one eye on these requirements, the designer of an RKE system must also juggle receiver sensitivity, carrier tolerance, and other technical parameters to achieve maximum transmission range within the constraints imposed by low cost and minimum supply current.

Further constraints include those defined by local regulations for short-range devices (such as FCC regulations in the US). The use of short-range devices does not require a license, but the products themselves are governed by laws and regulations that vary from country to country. For the US, the relevant document is the Code of Federal Regulations (CFR), Title 47, Part 15, which includes the 260-to-470-MHz band (Section 15.231) and the 902-to-928-MHz band (Section 15.249).[1]

The following provides some guidelines as to how the FCC regulations impose limits on an RKE design: Section 15.231 allows the device to transmit command or control signals, identification (ID) codes, and radio-control signals during emergencies, but not voice or video, toy-control signals, or continuous data. Transmission times must not exceed five seconds, and periodic transmissions of one second maximum at regular intervals are allowed only if the rate of such transmissions is less than one per hour.

Maximum field strength at three meters from the transmit antenna should be linearly proportional to the fundamental frequency (260 to 470 MHz), giving a range of 3750 to 12,500 μV/m. Bandwidths at points 20 dB down from the carrier should not exceed 0.25 percent of the center frequency, and spurious emissions should be attenuated by 20 dB of the fundamental-frequency signal.

First-generation RKE circuitry includes surface-acoustic-wave (SAW) source devices for generating an RF carrier in the transmitter and a local-oscillator (LO) frequency in the receiver. Unfortunately, the initial frequency uncertainty of a typical SAW device is at least ±100 kHz, and its frequency stability versus temperature is relatively poor. At the receiver, an intermediate-frequency (IF) pass band wide enough to admit the

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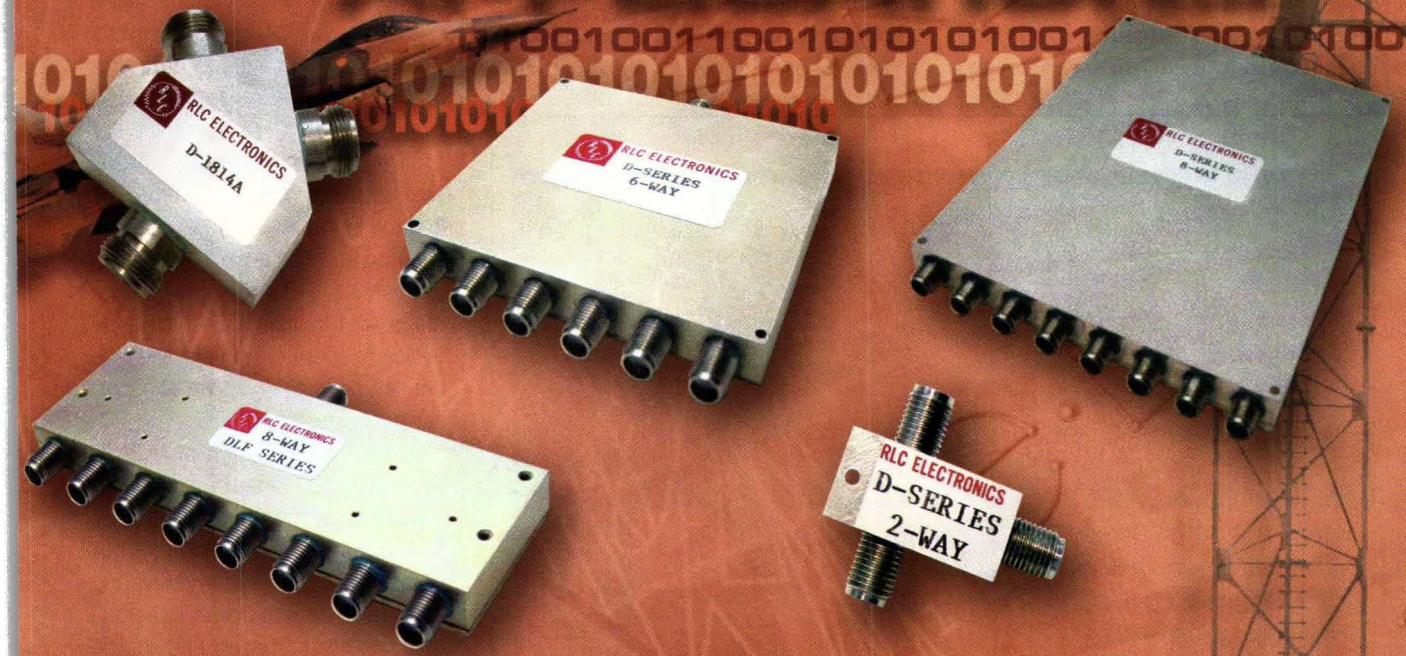

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carrier also admits excessive noise, which in turn limits the range at which the vehicle can respond to a key fob signal.

A current alternative to SAW devices is crystal-based phase-locked-loop (PLL) oscillator. (The transition to PLLs is encouraged by increasingly strict regulation of RF emissions, especially in Europe and Japan.) A crystal-based PLL transmitter costs only $0.50 (US) more than one with a SAW resonator, but is typically 10 times more accurate. The receiver can therefore have a narrower IF bandwidth, which in turn

RKE CMOS ICs

TARLTON FLEMING
Writer/Editor

Maxim Integrated Products (Sunnyvale, CA) is one of several manufacturers producing special-purpose integrated circuits (ICs) for the RKE market. For the key fob, it offers the world's smallest transmitter of its type: the 300-to-450-MHz MAX1472, which is supplied in a tiny 8-pin SOT23 package only 3 × 3 mm. Its supply voltage range (+2.1 to +3.6 VDC) enables the device to operate from a single lithium cell, drawing only 5 nA of supply current in standby mode.

During transmission of Manchester-encoded data, the MAX1472 supports data rates to 100 kb/s and draws between 3.0 and 5.5 mA of supply current while delivering –10 to +10 dBm of power to a 50-Ω load. Its crystal-based PLL source produces an accurate carrier frequency that enhances transmission range by allowing a tighter IF bandwidth in the receiver. To minimize power consumption, the internal oscillator starts quickly. It requires only 220 µs startup time following an enable signal.

The MAX1473 300-to-450-MHz superheterodyne amplitude-shift-keying (ASK) receiver, on the other hand, is ideal for use in vehicle RKE receivers. It offers –114 dBm sensitivity and 50 dB RF-image rejection in its fully differential internal mixer. The IC is optimized for either 315- or 433-MHz operation and operates on +3.3 or +5 VDC. The receiver includes a low-noise amplifier (LNA), crystal-based PLL LO, and a 10.7-MHz IF limiting amplifier with received-signal-strength-indicator (RSSI) circuitry. An internal data filter and data slicer provide the digital data output. As an alternative, the MAX1470 receiver is similar to the MAX1473 but optimized only for 315-MHz operation. It runs on a supply voltage of +3.0 to +3.6 VDC.

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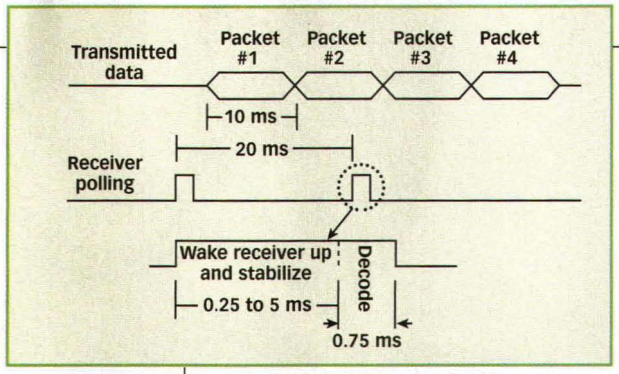
extends the transmission distance by raising the signal-to-noise ratio (SNR).

Earlier SAW devices were designed to operate at the midpoint of the 1.74-MHz-wide 433-MHz band (433.05 to 434.79 MHz) to ensure good system reliability even with expected process and temperature variations. Thus, the nominal carrier frequency for 433 MHz applications is now 433.92 MHz, and PLL crystals must be selected accordingly.



2. To monitor key fob transmissions, an RKE receiver must allocate time to wake up and stabilize before decoding the incoming signal.

Modern receiver and transmitter ICs feature integrated PLL circuitry requiring only an external crystal resonator for effective RKE signal generation **(see sidebar)**. The MAX1470 PLL from Maxim Integrated Products, for example, includes a divide-by-64 block and a 10.7-MHz IF with low-side injection. (The chip can operate at 433.92 MHz, but its image-rejection capability is optimized for 315 MHz.) The required crystal frequency for 315 MHz operation (in MHz) is $f_{XTAL} = (f_{RF} - 10.7)/64 = 4.7547$. The IC is designed for use with a crystal that is specified to oscillate at 315 MHz when loaded with the 5-pF capacitance presented by chip terminals XTAL1 and XTAL2. For details on how to trim the crystal frequency, refer to Application Note 1017 available at the company's website.[2]

Because maximum battery life is important, RKE systems employ various techniques to minimize operating current and "on time." The voltage-controlled oscillator (VCO) in the receiver PLL offers a good example of this attention to detail. The receiver must check almost constantly to avoid missing a demand for entry to the vehicle, and to save power it attempts to shut down as often as possible, even during the brief intervals between checks.

A key fob transmitter usually issues four 10-ms data streams in succession (about 40 ms total) to ensure that the receiver captures at least one of them. The receiver performs a polling operation every 20 ms, seeking to decode at least two data streams as a margin against timing errors and noise. About 0.75 ms of decoding time (enough for receiving 7 or 8 data bits) is required

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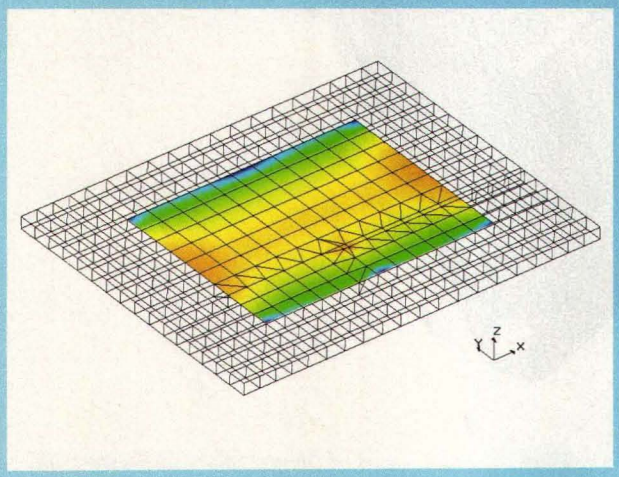
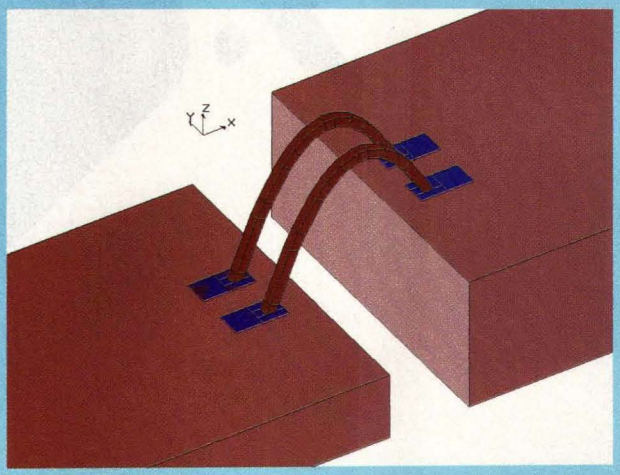
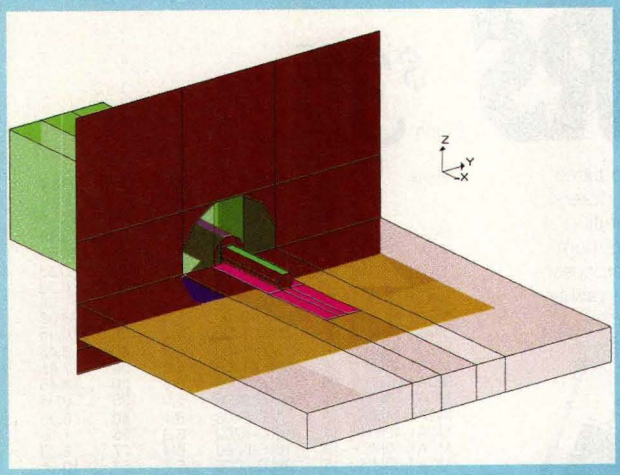


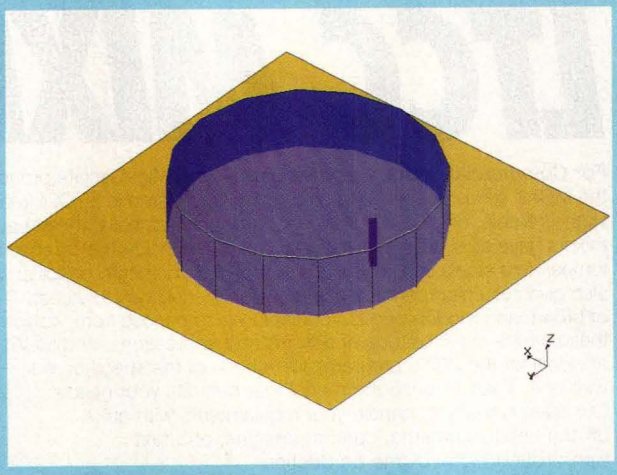
A patch antenna with finite size substrate



Wire bonds in inhomogeneous dielectrics



A coaxial to microstrip transition



A dielectric antenna

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to determine whether the data is of interest.

In addition to decoding time, the polling operation must first allow time for the receiver circuits to "wake up" and stabilize. Most amplifier circuits can wake up quickly, but the VCO's crystal is an electromechanical component that requires time to begin oscillating and more time to stabilize at the desired frequency. Conventional superheterodyne receivers require 2 to 5 ms for that purpose. The MAX1470's VCO does it in only 0.25 ms by supplying just enough power to maintain vibration in the crystal. Thus, the IC detects key fob transmissions by waking up for only 1 ms (0.75 ms for decoding plus 0.25 ms for stabilizing) during every 20 ms **(Fig. 2)**. The fast-wakeup MAX1470 also operates on +3.3 VDC instead of +5 VDC, for a net energy savings that extends battery life (with respect to conventional superheterodyne receivers) by a factor of four or five.

RKE is strictly a short-range technology (up to 20 m for active systems or 1 to 2 m for passive RKE systems), but ensuring even a short transmission distance on low power and a low-cost design budget can be challenging for the RF circuitry. For simplicity, the transmit and receive antennas consist of a circular or rectangular loop of copper trace on a small PCB, with a simple inductive-capacitive (LC) network to match the antenna impedance to the transmit or receive chip.[3]

The low transmit power imposed by FCC regulations, small battery capacity, and uncertainty in orientation of the transmit antenna demands maximum sensitivity at the RKE receiver chip. One way to enhance receiver sensitivity is to add an external low-noise amplifier **(Fig. 3)**, but the restriction in dynamic range associated with that approach might be unacceptable in a given application. For example, consider the following analysis based on the MAX1470 superheterodyne receiver.

A receiver's sensitivity depends on its noise figure, the minimum SNR required for detection of the carrier modulation, and the thermal noise in the system:

$$S = NF + n_0 + S/N \quad (1)$$

where:

S = the minimum required signal level (in dBm);

NF = the receiver noise figure (in dB);

n_0 = the receiver's thermal noise power in dBm; and

S/N = the output SNR (in dB) required for required for adequate signal detection (usually based on the acceptable bit-error rate).

For simplicity, an SNR of 5 dB is estimated, based on an assumption of Manchester-encoded data. By definition,

$$n_0 = 10\log_{10}(kTB/1 \times 10^{-3}),$$

where:

k = Boltzmann's constant (1.38×10^{-23});

B = the system noise bandwidth; and

T = the temperature (in °K).

At room temperature (T = 290°K) over a 1-Hz bandwidth n_0 = –174 dBm/Hz. Over a 300-kHz IF bandwidth, n_0 = –119 dBm.

Assuming the system sensitivity (S) is –109 dBm, Eq. 1 can be used to calculate a noise figure (NF) of 5 dB. The relationship between noise figure (NF) and noise factor (F) is $(NF)_{dB} = 10\log F$, where $F = 10(NF_{dB}/10)$. Thus, F = 3.162. For a cascade of several two-port devices, the noise factor is:

$$F_{total} = F1 + (F2 - 1)/G1 + (F3 - 1)/(G1 \times G2) + \ldots \quad (2)$$

where:

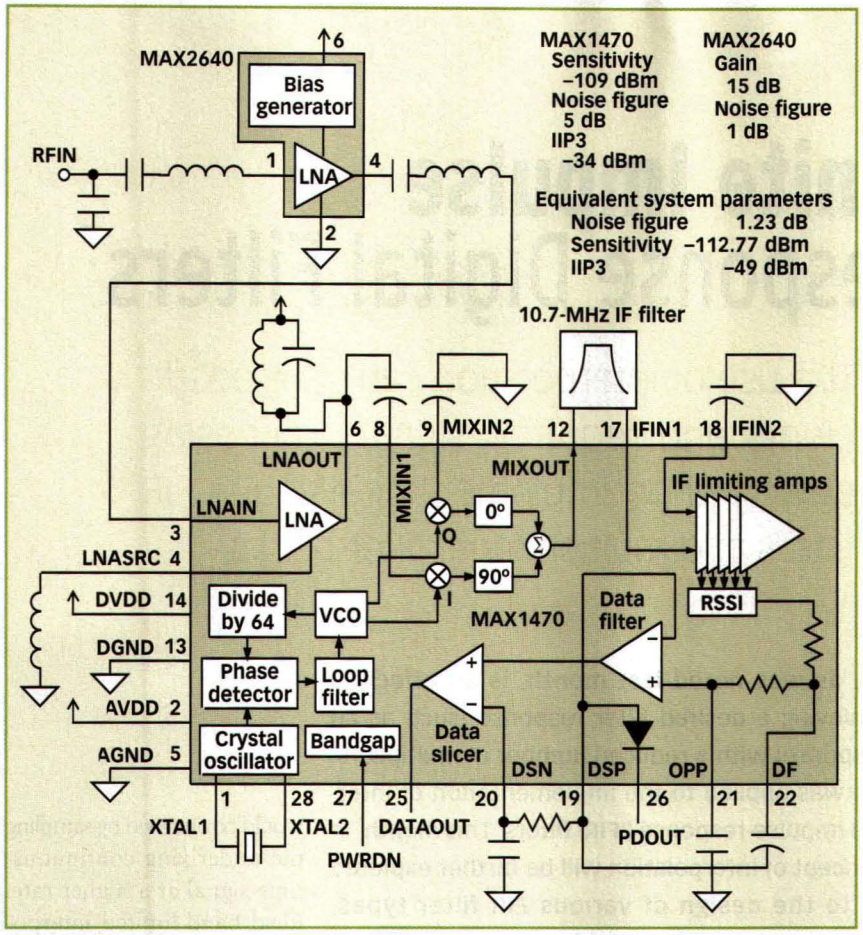
F1, F2, F3 = the noise factors of system stages one, two, and three, respectively, and

G1, G2, G3 = the gain (numerical voltage) of system stages one, two, and three, respectively.

Equation 2 makes it possible to calculate the new noise factor after adding an external LNA to the system. For the MAX2640 LNA from Maxim, NF = 1 dB and gain = 15 dB (i.e., F1 = 1.26 and G1 = 31.62). The noise factor for the original system was 3.162, so $F_{Total} = 1.327$, which is 1.23 dB. Substituting this value into Eq. 1 yields:

$$S = 1.23 - 119 + 5 = -112.77 \text{ dB}$$

Assuming that the original sensitivity was –109 dB, only 3.77 dB has been gained by the addition of the LNA. Note the effect on dynamic range as indicated by the third-order intercept point (IIP3). The MAX1470 has an internal LNA gain of 16 dB and an internal mixer IIP3 of –18 dBm, for an overall



3. Adding an external LNA (a model MAX2640) increases receiver sensitivity, but lowers the third-order intercept point and system dynamic range.

IIP3 of –34 dBm. Adding the external LNA with its gain of 15 dB lowers this number to –49 dBm. Thus, the addition of an external LNA improved sensitivity by almost 4 dB, but reduces the system dynamic range by 15 dB! For a given application, such a trade-off must be considered.

Beyond simplex systems, the next RKE advancement involved two-way, half-duplex systems which first appeared as the "passive RKE" already available in some high-end automobiles. The vehicle's transmitter is continually polling to detect the proximity of an operator's key fob. Within range (1 to 2 m), the key fob and vehicle establish two-way communications and open the door. Current two-way systems include the usual acknowledgment functions in addition to a remote-start function that allows an operator to start the engine from a distance.

Future developments may also include the technology for tire-pressure sensing (TPS). Like passive RKE, TPS is available at this time only for some trucks and luxury automobiles. TPS systems have much in common with RKE. Circuitry very similar to that of an RKE key fob resides in the valve stem of each tire, along with a sensor for tire pressure and temperature. Regular transmissions from each tire to a receiver in the vehicle (very similar to an RKE receiver) then provide the driver with an early warning of any problem developing with the tires. TPS and RKE have so much in common (short range, simple modulation, need to conserve power, etc.), that future systems will probably save costs by sharing and consolidating circuit functions. MRF

REFERENCES

1. For more information on the Federal Communications Commission and RF devices, go to the website at www.access.gpo.gov/nara/cfr/waisidx_01/47cfr15_01.html.
2. Application Note 1017, Maxim Integrated Products, Sunnyvale, CA, www.maxim-ic.com.
3. Application Note 1830, Maxim Integrated Products, Sunnyvale, CA, www.maxim-ic.com.

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The use of interpolation can save coefficients and required signal-processing power when implementing Nyquist FIR filters and multichannel filter banks,

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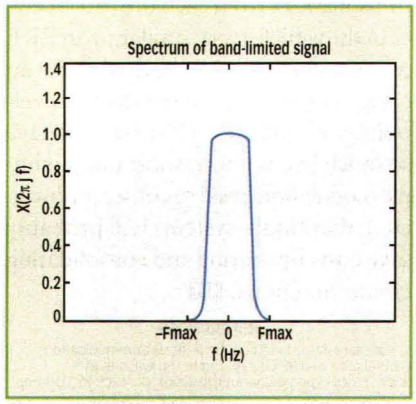
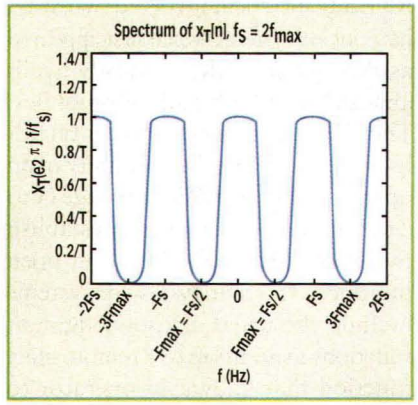
Interpolation, as was noted last month, is an effective means of achieving a desired filter response (such as an equiripple response) with a reduced number of multipliers. The approach was applied to the implementation of interpolation finite-impulse-response (IFIR) filters. This month, in Part 3, the concept of interpolation will be further explored and applied to the design of various FIR filter types, including Nyquist filters and more complex filter banks.

In the context of multirate signal processing, interpolation usually refers to band-limited interpolation. Band-limited interpolation is based on the notion of an underlying band-limited continuous-time signal that is being sampled. Ideal band-limited interpolation will take a digital (sampled) signal and produced an interpolated signal that will be identical to the signal that would be obtained by sampling the underlying continuous-time signal at a higher rate. Ideal band-limited interpolation can be accomplished by means of upsampling and using an ideal low-pass filter. Especially interesting is a time-domain interpretation of the ideal interpolator, which leads naturally to polyphase implementations.

The key concept of band-limited-interpolation is that a signal to be interpolated is a sampled version of a band-limited continuous time signal. Denote the continuous-time signal by $x_c(t)$ and suppose its spectrum is zero for all |f|



25. This spectrum shows a band-limited continuous-time signal.



26, This spectrum shows a sampled signal with sampling rate, $f_s = 2f_{max}$.

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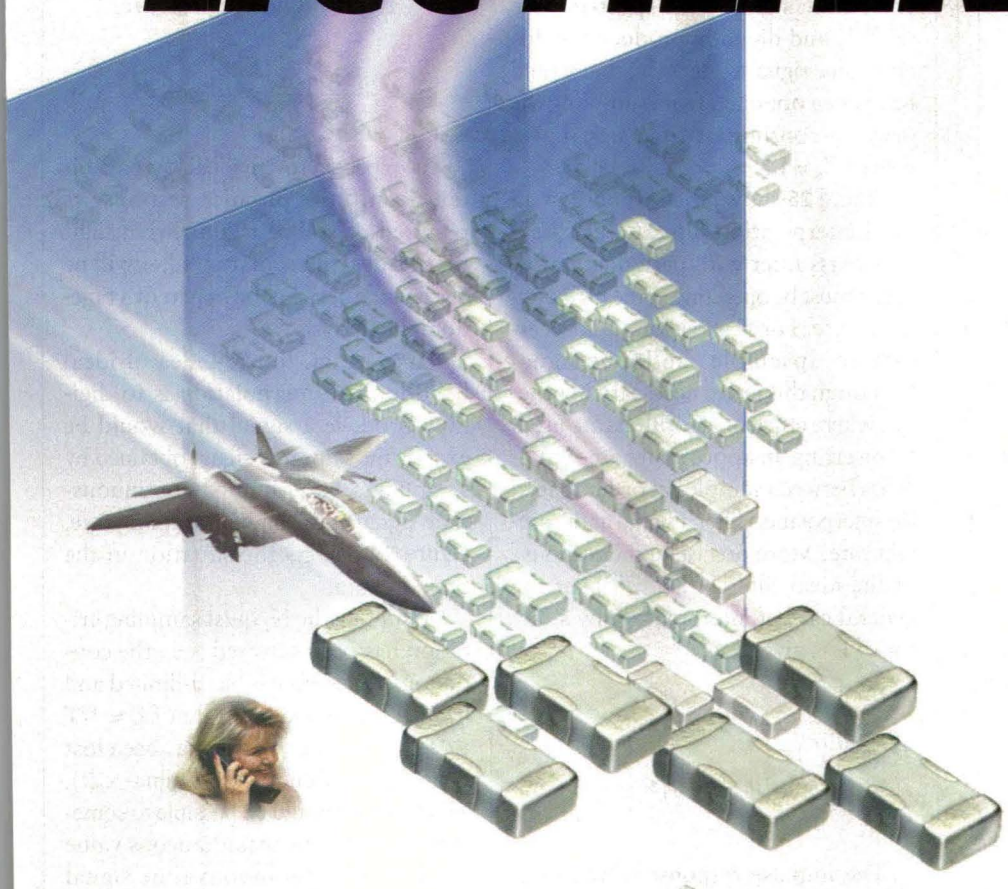

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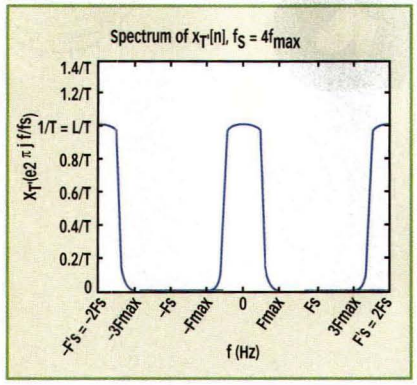


27. This spectrum shows a sampled signal with sampling rate of $f'_s = 4f_{max}$.

$> f_{max}$. **Figure 25** shows the frequency spectrum $X(2\pi jf)$. If the signal is sampled at a sampling frequency $f_s = 2f_{max}$, the following signal results:

$$x_T[n] = \{ x_c(nT) \},\ T = 1/f_s$$

Figure 26 shows its spectrum, $xT(e^{2\pi jf/fs})$. Now suppose the continuous-time signal was sampled at a rate of $f's = Lf_s = 2Lf_{max}$. The sampled signal at the higher rate can be represented as:

$$x_{T'}[m] = \{xc(mT')\},\ T' = 1/f's = T/L,$$

where:

$$m = Ln + k,\ k = 0 \ldots (L-1)$$

will have a spectrum $x_{T'}(e^{2\pi jf/fs})$ as shown in **Fig. 27** for the case L = 2.

The job of the ideal interpolation filter should now be clear from the frequency-domain standpoint. Take the discrete-time signal with spectrum $xT(e^{2\pi jf/fs})$ and digitally produce the discrete-time signal $x_{T'}(e^{2\pi jf/f's})$ that would have been obtained from sampling the original continuous-time signal at a rate of $f'_s = Lf'_s$.

Figure 28 shows the response of an ideal interpolation filter. Clearly, it is a lowpass filter with periodicity of f'_s, i.e. it must be operating at the high sampling rate. For this reason, it is necessary to upsample the input signal (although this is not necessary in practice where efficient algorithms are used) by inserting an appropriate amount of zeros between samples in order to feed the interpolation filter a signal at the correct rate. More precisely, the response of the ideal filter $H_D(e^{2\pi jf/f's})$ for the general case of interpolation by a factor of L is given by:

$$H_D\left(e^{2\pi jf/f'_s}\right) = \begin{cases} L, & |f| \le \frac{f_s}{2} \\ 0, & \frac{f_s}{2} < |f| \le \frac{f'_s}{2} \end{cases} \qquad (3)$$

The impulse response of the ideal interpolation filter can be found from the inverse DTFT (ref. 1):

$$h_D[m] = \frac{L}{f'_s}\frac{\sin(\pi f_s T' m)}{\pi T' m} \qquad (4)$$

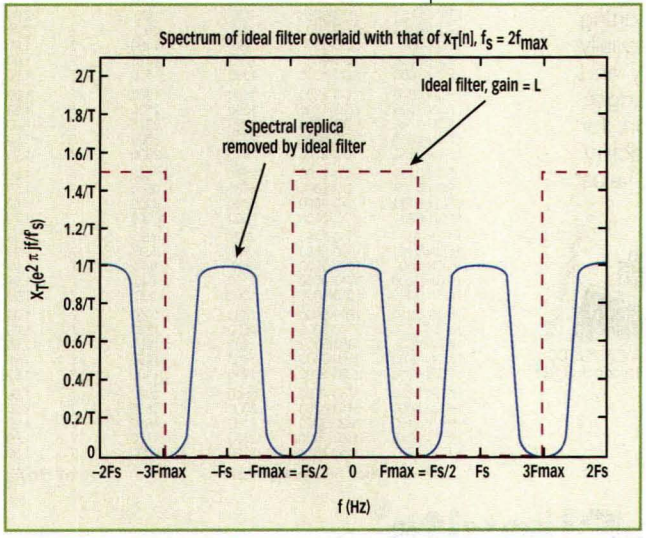
Using the fact that $Lf'_s = 1/T'$ and $T' = 1/Lf_s$, the following results:

$$h_D[m] = \frac{\sin(\pi m/L)}{\pi m/L},\ -\infty < m < \infty \qquad (5)$$

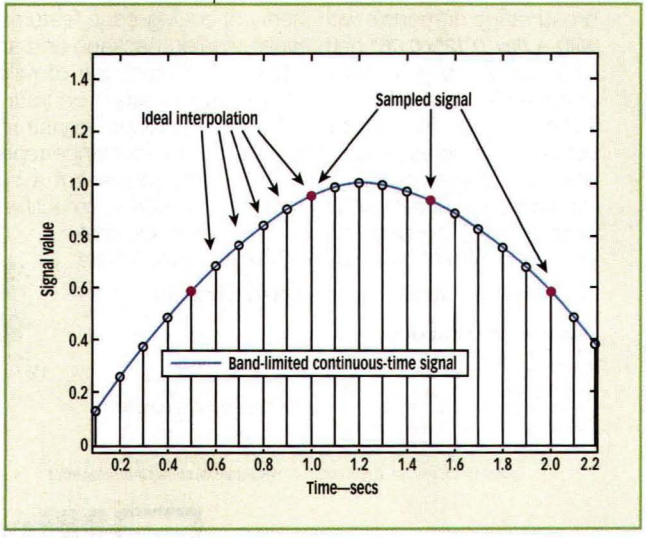
As expected for an ideal lowpass filter, an infinite impulse response is required to realize it. Further insight for the ideal interpolation filter will be given in the next section as part of a time-domain analysis.

Once again, the key idea of ideal band-limited interpolation is to digitally produce a signal that would be exactly the same as a signal obtained by sampling a band-limited continuous-time signal at the higher sampling rate. **Figure 29** depicts the situation in the time domain.

Assuming the Nyquist sampling criterion has been satisfied (i.e., the continuous-time signal is band-limited and has been sampled at a rate of $f_s = 1/T \le 2f_{max}$), no information has been lost from the continuous-time signal $x_c(t)$. Therefore, it should be possible to somehow recreate any instantaneous value $x_c(t_0)$ of the continuous-time signal from the sampled signal $x_T[n]$. Looking at Fig. 29, it can be seen that the job of the fivefold interpolator is to take every input sample $x_T[n]$ and produce five output samples $\{x_{T'}[m]\}$, m = 5n + k, k = 0...4 as follows (note that T = 0.5 and



28. The response of an ideal interpolation filter is overlaid here with the spectrum of a signal sampled at a rate of $f_s = 2f_{max}$.



29. This plot of an ideal band-limited interpolation is depicted in the time domain.

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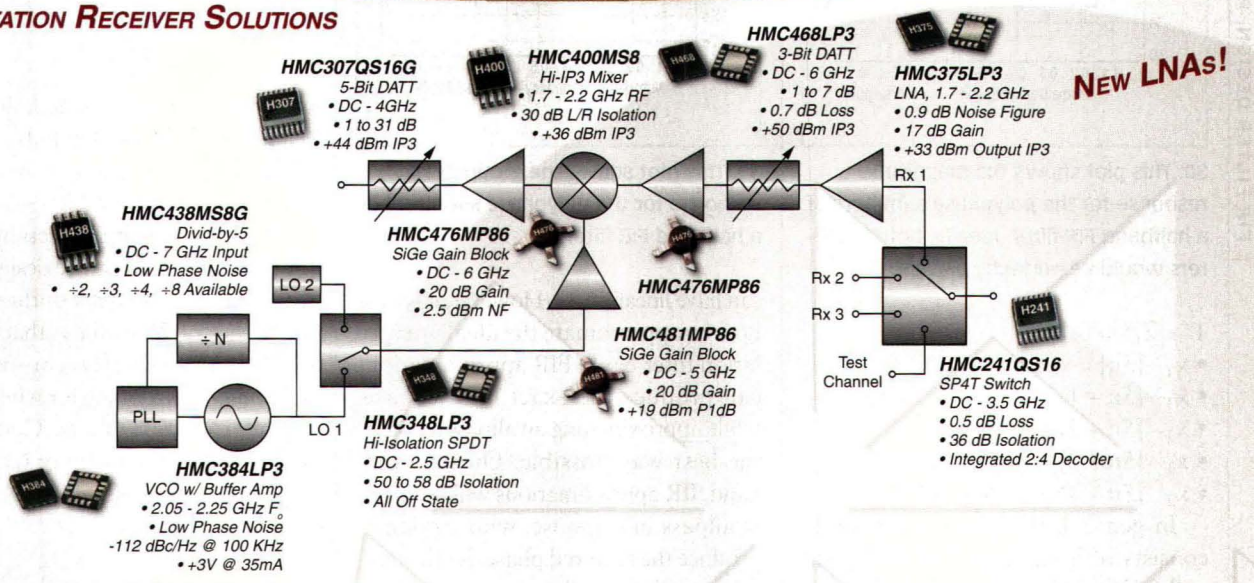

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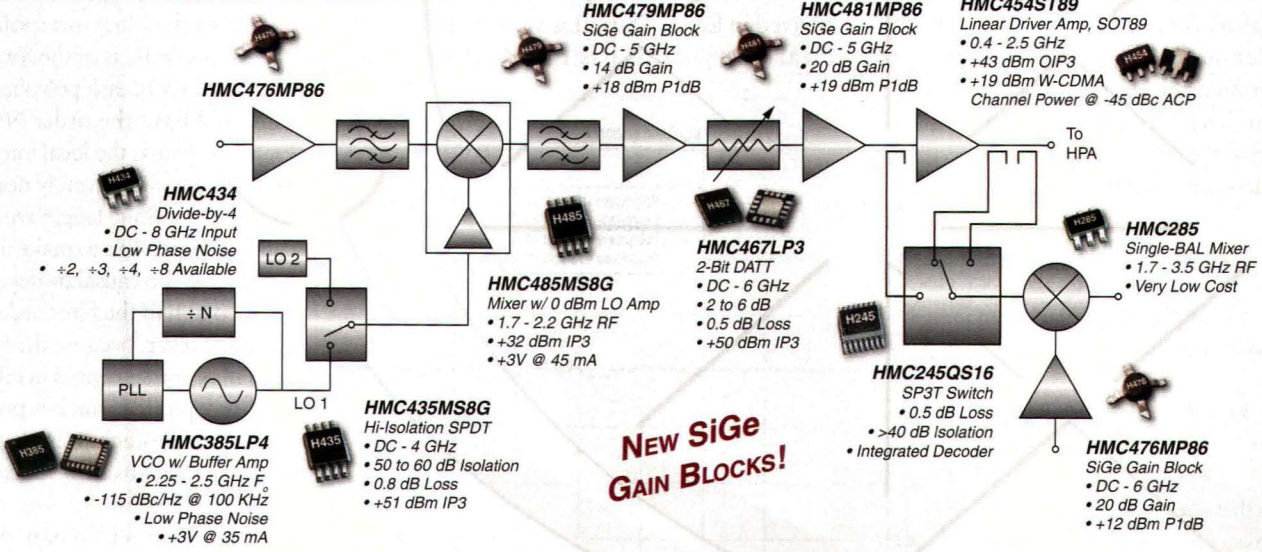

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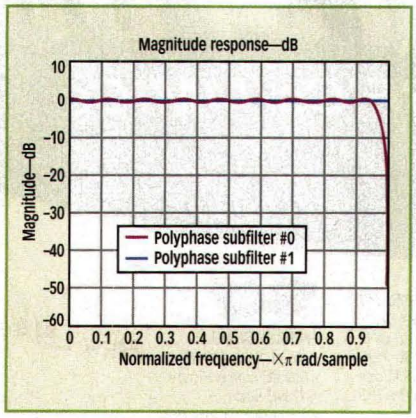
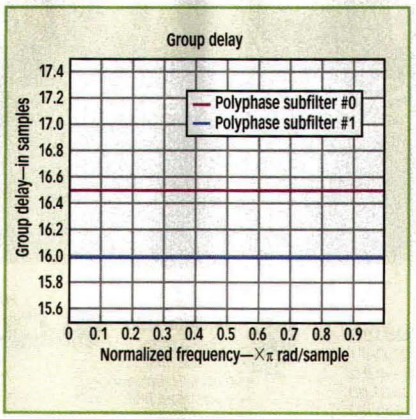


30. This plot shows the magnitude response for the polyphase subfilters of a halfband FIR filter. Ideally, both subfilters would be perfectly allpass.



31. This plot shows the group-delay response for the polyphase subfilters of a halfband FIR filter.

$T' = T/5 = 0.1$):.

- $x_{T'}\ [5n] = x_T\ [n]$
- $x_{T'}\ [5n + 1] = x_T\ [n + 1/5\ ]$
- $x_{T'}\ [5n + 2] = x_T\ [n + 2/5\ ]$
- $x_{T'}\ [5n + 3] = x_T\ [n + 3/5\ ]$
- $x_{T'}\ [5n + 4] = x_T\ [n + 4/5\ ]$

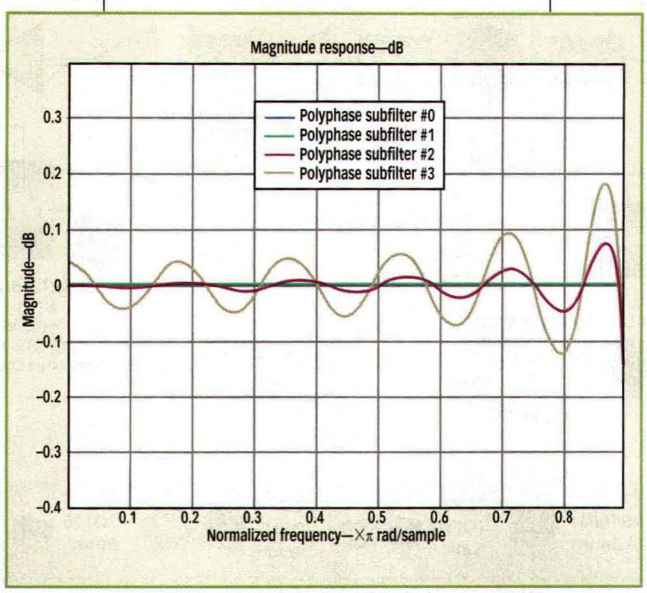
In general, the ideal interpolator consists of a bank of L filters which will *fractionally advance* the input signal by a factor of k/L, k = 0...(L – 1). The outputs of the filters are then interleaved (i.e., only one filter needs to operate per high rate output sample) to produce the interpolated signal. The L filters that comprise the filter bank are the fractional advance filters $H_k(z)$:

$$H_k(z) = z^{k/L},\ k = 0...(L-1)$$

Evaluating this on the unit circle results in

$$H_k(e^{j\omega}) = e^{j\omega k/L},\ k = 0...(L-1)$$

so that each filter $H_k(e^{j\omega})$ is allpass, i.e., $|H_k(e^{j\omega})| = 1$ and has linear phase, $\arg\{H_k(e^{j\omega})\} = \omega k/L$.Herein lies the impossibility of designing these filters. They cannot be designed as FIR filters because no FIR filter can be allpass (except for a pure delay). They cannot be designed as IIR filters, because no stable IIR filter can have linear phase. However, it is clear how to approximate the ideal interpolation filter bank. FIR approximations can produce the exact linear phase, while approximating an allpass response the best way possible. On the other hand, IIR approximations will be exactly allpass in response, while trying to produce the required phase. It is insightful to realize that the filters comprising the filter bank are the polyphase components of the ideal interpolation filter derived in Eq. 5. Thus, this view of the ideal interpolator has the efficient polyphase structure "built-in."



32. This plot shows the magnitude response for the polyphase subfilters of a Nyquist FIR filter designed with the window method. The polyphase subfilters better approximate allpass filters than a comparable equiripple design for the majority of the frequency band.

Indeed, the impulse response of each fractional advance filter in the filter bank is given by the inverse DTFT,

$$h_k[n] = \frac{1}{2\pi}\int_{-\pi}^{\pi} e^{j\omega k/L} e^{j\omega n} d\omega = \frac{\sin\left(\pi \dfrac{Ln+k}{L}\right)}{\pi \dfrac{Ln+k}{L}}$$

which corresponds to the L decimated sequences of the ideal impulse response by again writing uniquely $m = L_n + k$, $k = 0 \ldots (L - 1)$ in Eq. 5.

While interpolation filters are simply lowpass filters that can be designed with the various techniques outlined previously, the polyphase filters that compose the ideal interpolation filter give some insight on things to be looking for when designing interpolation filters. Consider an interpolation by a factor of L. The ideal L polyphase filters will have a group-delay given by:

$$-k/L,\ k = 0 \ldots (L-1)$$

For simplicity, consider an FIR approximation to the ideal interpolation filter where the order is of the form N = 2LM. Them each polyphase filter will have the order N/L = 2M. Note that the ideal interpolation filter is infinitely noncausal. After finite length truncation, it is possible to make the approximation causal by delaying by one-half the filter order, N/2. However, because the filter will be implemented in efficient polyphase form, it is possible to make each polyphase component causal by delaying it by M samples.

The delay will mean the introduction of a phase component in the response of each polyphase component. So that instead of approximating the ideal fractional advance $e^{j\omega k/L}$ the polypase components will approximate $e^{j\omega(k/L - M)}$. Consequently, the group delay will have the form

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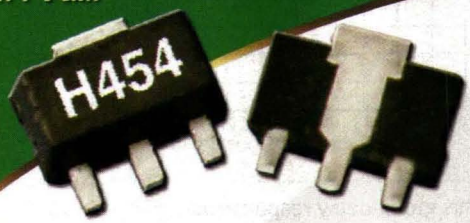

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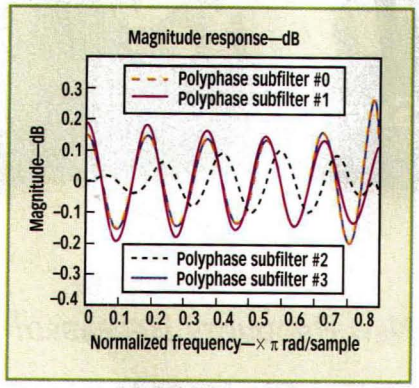
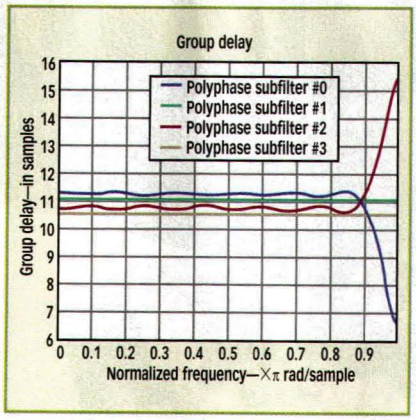
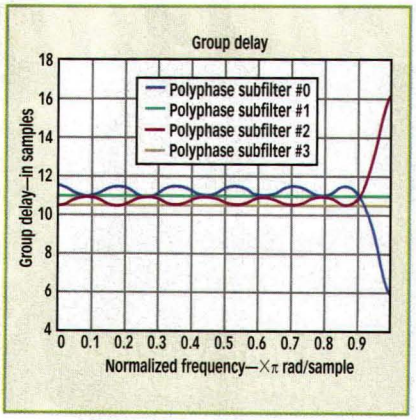


33. This plot shows the magnitude response for the polyphase subfilters of a optimal equiripple lowpass FIR filter. None of the subfilters behaves as a perfect allpass, an indication that this is not a Nyquist filter.



34. This group-delay response was generated for the polyphase subfilters of a Nyquist FIR filter of order 90 and L = 4.



35. This group-delay response was generated for the polyphase subfilters of a conventional equiripple lowpass design that could be used for interpolation with L = 4.

$$-\frac{d\phi(\omega)}{d\omega} = -\frac{d\omega(k/L - M)}{d\omega} = M - k/L$$

A problem that arises is that even though the FIR approximation to the ideal interpolation filter is symmetric and thus has linear phase, the polyphase components are not necessarily symmetric and thus will not necessarily have exact linear phase. However, for each non symmetric polyphase filter, there is a mirror image polyphase filter which will have the exact same magnitude response with a mirror image group delay that will compensate any phase distortion.

When we analyzed the behavior of the ideal interpolation filter in the time-domain, we saw that for every input sample, L samples are produced including one that is exactly the same as the input sample. This exact copy is "produced" by the polyphase filter that has allpass magnitude and zero phase (i.e., the case where k = 0). In practice, this is the only polyphase filter that can be designed exactly, albeit with a group delay of M rather than 0. Roughly speaking, a Nyquist filter is one for which one of its polyphase components is a pure delay and thus leaves the input signal unchanged (except for a possible delay). When designing an interpolation filter, it is desirable for it to be a Nyquist filter since this will ensure that even a non-ideal filter will allow the input samples to pass through unchanged. It can also be computationally advantageous since one of the polyphase subfilters will have no multipliers.

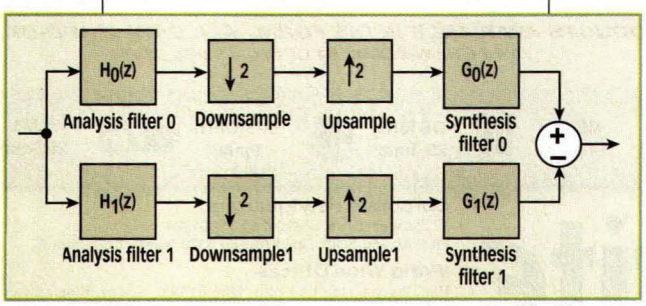
Nyquist filters are also called Lth-band filters because the passband of their magnitude response occupies roughly 1/L of the Nyquist interval. In the special case of an interpolation by a factor of 2, the filters are known as halfband filters. Halfband filters are commonly used when interpolating (or decimating) by a factor of 2. The cutoff frequency for a halfband filter is always 0.5π. Moreover, the passband and stopband ripples are identical, limiting the degrees of freedom in the design. The function "firhalfband" designs FIR halfband filters. The specifications set still follows the triangle metaphor shown in Fig. 2, taking into account the limitations just described. The following three function calls design three equiripple linear-phase halfband filters using a different pair of specifications in each case from the three available-order (N), transition-width (TW), and peak passband/stopband ripple (R):

```
b1 = firhalfband(102, .47);
%N and TW
b2 = firhalfband(102, .01, 'dev');
%N and TW
b3 = firhalfband('minorder', .47,
.01);     %TW and R
```

It is possible to analyze how the design compares to the ideal interpolation filter by creating an FIR interpolator object and looking at its polyphase subfilters, for example, using the third filter, b3,

```
h = mfilt.firinterp(2,2*b3);
polyphase(h)
```

Figures 30 and 31 show the magnitude and group-delay responses for the polyphase components of this filter. Note that M = N/2L is 16.5 in this case, so that the group delays are exactly M – k/L, k = 0, 1. The only deviation from an ideal filter (ignoring an overall delay of *M* samples) comes from the fact that one of the polyphase subfilters is not a perfectly allpass response.



36. This block diagram shows the two-channel FIR subband filter bank.

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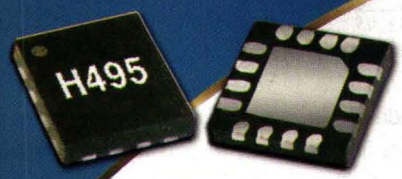

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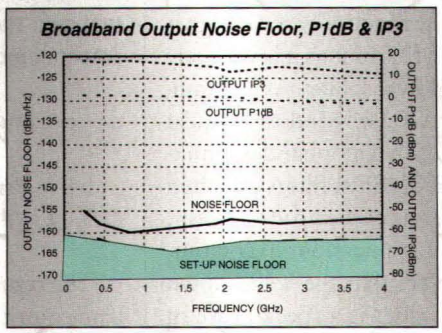

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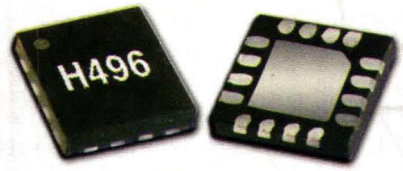

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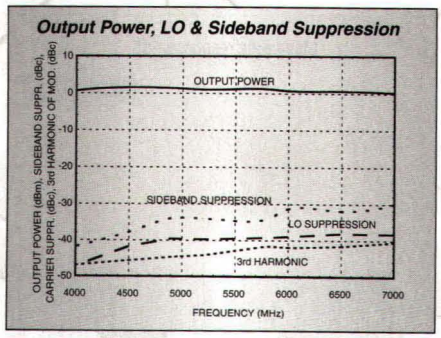

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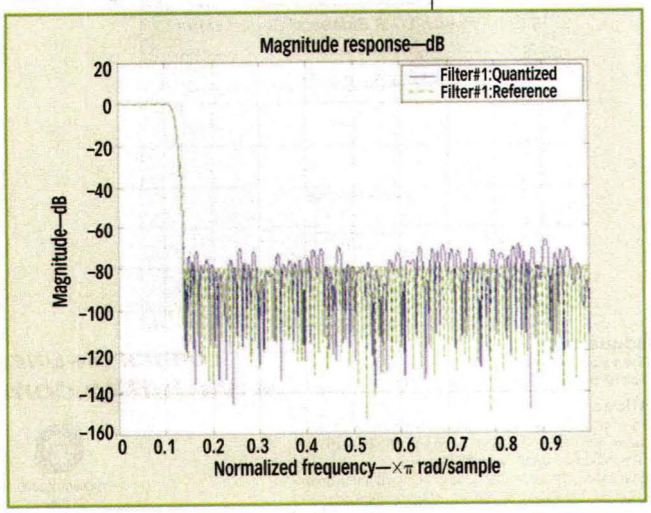
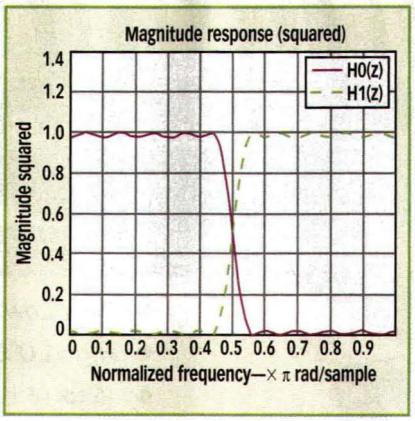
Nyquist filters are characterized in the time-domain by their impulse response being exactly equal to zero every L samples (except the exact middle sample of the impulse response). This is precisely why we get a polyphase sub filter that is a perfect allpass delay and allows the samples to be interpolated to pass through the filter unchanged. Designing a filter that is both a lowpass and simultaneously satisfies the just mentioned time-domain characteristic is not a trivial task except for the case of window-based designs.[13, 14]

Nevertheless, the advantage of conventional optimal equiripple designs over a Nyquist window-based design is not as clear in this case as it is with any conventional lowpass filter. As an example, consider a Kaiser window Nyquist filter design with a stopband attenuation of 40 dB. Nyquist filters are often designed in terms of their rolloff factor, ρ, due to their applications in communications. [Note that the well-known raised-cosine filter is a special case of a Nyquist filter. In fact, the same reason that raised-cosine filters are common, i.e., to achieve zero intersymbol-interference (ISI) with a nonideal filter, is why they are able to interpolate without affecting the input samples—namely the fact that the impulse response becomes zero exactly at the right time.] The rolloff factor is related to transition-width (TW) supply by $TW = \rho\pi/L$. In this example, $\rho = 0.1$ and $L = 4$; thus, the transition width is 0.0025π:



38. This plot shows the magnitude response of the filter quantized with [16,15] format.



37. This plot shows the magnitude-squared responses of the analysis filters in an FIR perfect reconstruction filter bank. The two filters are power-complementary.

```
b1 = firnyquist('minorder', 4, .1, .01);  L = 4
```

The resulting filter is of 90th order. If an equiripple filter of the same order and same attenuation is designed, it will result in a filter with a smaller transition width, but on that does not satisfy the time-domain requirement:

```
b2 = firceqiup(90, .25, [.01 .01]);
```

Figure 32 shows the magnitude responses of the polyphase subfilters for the Nyquist window-based design. For comparison, **Fig. 33** shows the magnitude responses for the optimal equiripple design. Note the better approximation to allpass filters in the Nyquist design compared to the equiripple design (albeit for a slightly smaller interval, which is the trade-off). Similarly, if we compare the group-delay response of the polyphase subfilters, the Nyquist design once again better approximates the ideal constant group-delay as compared to the equiripple design. **Figure 34** shows the group-delay responses for the polyphase subfilters of the Nyquist design. **Figure 35** shows the group-delay responses for the polyphase subfilters of the equiripple design.

Figure 36 shows a two-channel filter bank. Filters $H_0(z)$ and $H_1(z)$ are called the analysis filters while $G_0(z)$ and $G_1(z)$ are the synthesis filters. The filter bank is called perfect reconstruction if the end-to-end system acts as a delay, i.e., if the output signal is simply a delayed (and possibly scaled) version of the input.

It is well known[10, 15, 16] that perfect reconstruction can be achieved if

$$0.5G_0(z)H_0(-z) + 0.5G_1(z)H_1(-z) = 0$$

and

$$0.5G_0(z)H_0(z) + 0.5G_1(z)H_1(-z) = z^{-k}$$

Starting with a prototype lowpass filter H(z) of odd order N, the following selection for the filters results in perfect reconstruction using solely FIR filters[15]:

$$H_0(z) = H(z)$$
$$H_1(z) = z^{-N}H_0(-z^{-1})$$
$$G_0(z) = 2z^{-N}H_0(z^{-1})$$
$$G_1(z) = 2z^{-N}H_1(z^{-1})$$

This type of perfect reconstruction filter bank is called an orthogonal filter bank or a power-symmetric filter bank.[15]

The function "firpr2chfb" designs equiripple FIR filters H0(z), H1(z), G0(z), G1(z) such that the filter bank is an orthogonal perfect reconstruction filter bank. The parameters to specify are simply the filter order *N* and the passband-edge frequency *wp*. A prototype lowpass filter is designed from which the four required filters are obtained.

For example,

```
[h0,h1,g0,g1] = firpr2chfb(99,.45);
```

Alternatively, the peak stopband ripple can be specified. As usual, we can obtain minimum-order designs by

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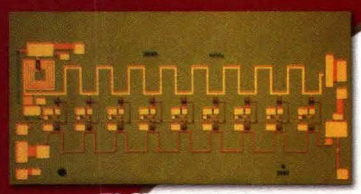

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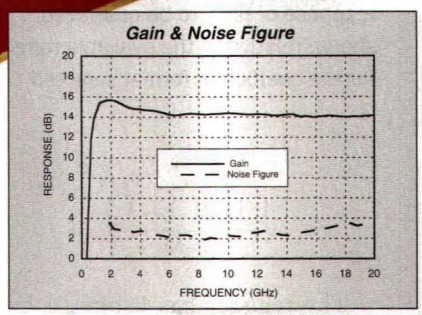

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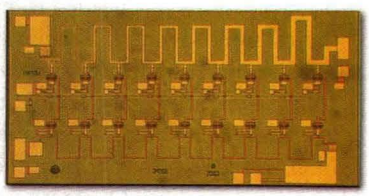

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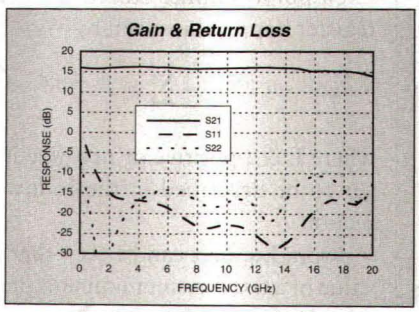

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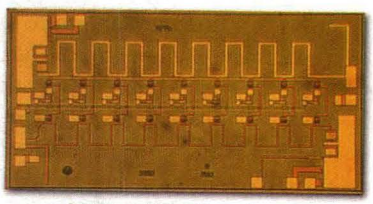
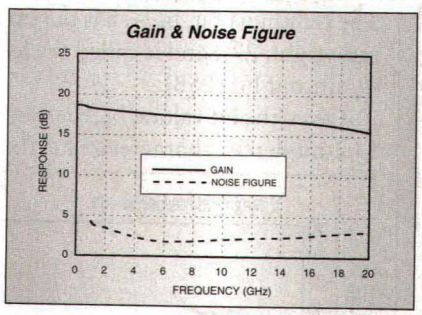


50 Ohm I/O Driver Amp



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specifying both the passband-edge frequency and the peak stopband ripple. In all cases, the design specifications apply to the prototype filter H(z),

```
[h0,h1,g0,g1] = firpr2chfb(99, le-3, 'dev');
[h0,h1,g0,g1] = firpr2chfb('minorder', .45, le-3);
```

The power-symmetric term is used because for these designs, the following holds:

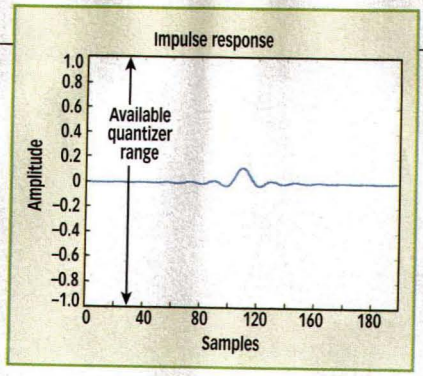
$$\left|H_0\left(e^{j\omega}\right)\right|^2 + \left|H_1\left(e^{j\omega}\right)\right|^2 = 1,\ \forall\omega$$

It is possible to look at the magnitude-squared responses of $H_0(z)$, and $H_1(z)$ using the "fvtool" function. **Figure 37** shows the magnitude-squared responses for N = 19 and $\omega_p = 0.45\pi$. Notice how where one filter's ripple rises the other filter's ripple declines to add up to one.

Increasing the filter order (and possibly the passband-edge frequency) improves the lowpass/highpass separation provided by the analysis filters but does not have an effect on the perfect reconstruction characteristic of the overall system.

Several factors have to be taken into account when implementing an FIR filter using fixed-point arithmetic. For one thing, the coefficients have to be quantized from double-precision floating point in which they are designed into fixed-point representation with usually a smaller number of bits. It is necessary to make sure to make the most of the limited number of bits available. Furthermore, performing the arithmetic in fixed-point will introduce further quantization errors when actually filtering with the quantized coefficients. These quantization errors should be minimized as much as the hardware at hand allows.

The Filter Design Toolbox of MATLAB employs certain notation to represent fixed-point numbers. Consider a register used to store a fixed-point number,



39. This plot shows the impulse response of filter to be quantized shown relative to the available range for the coefficient format selected.

$$\underbrace{(\,)(\,)(\,)\ldots(\,)}_{B-bits}\quad \text{over } b_0\ b_1\ b_2\ \ldots\ b_{B-1}$$

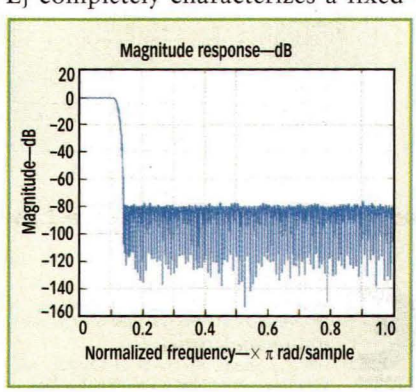
The register has B bits (it has a wordlength of B), the value of the kth bit is given by b_k which can obviously be only 0 or 1. A two's complement fixed-point number stored in such a register has a value given by

$$value = -b_0 2^{B-L-1} + \sum_{k=1}^{B-1} b_k 2^{B-L-k-1} \qquad (6)$$

where L is a positive or negative integer to be described in the following section.

From Eq. 6, it can be seen that the value of a fixed-point number is determined by assigning weights of the form 2 – m to each bit. The leftmost bit, b_0, has the largest weight, 2^{B-L-1}, and is called the most-significant bit (MSB). The rightmost bit, b_{B-1}, has the smallest weight, 2^{-L}, and is called the least-significant bit (LSB).

Given the bit values, b_k, the pair {B, L} completely characterizes a fixed-point number, i.e., suffices in determining the value that the bits represent. Such a pair is called the format of a given quantity, and is stored in a two-element vector, [B, L].



40. This plot shows the magnitude response of the filter quantized with [16,18] format.

Consider the following filter

```
b = gremez('minorder',[0 .11 .14 1],...
[1 1 0 0],[.01 .0001]);
```

The filter has an attenuation of 80 dB and a its largest coefficient is 0.1206.

The first thing to do is check if there are enough bits available to represent the coefficients and provide the required dynamic range. A good rule of thumb[17] is to assume 5 dB/b for the dynamic range (note that the usual 6 dB/b rule does not apply because the quantization error for the filter coefficients tends to be correlated, especially at the extremes of the impulse response). In this example, at least 16 b is needed to provide the required dynamic range (80 dB). But it is not sufficient to simply apply 16 b to the design. For example, the following code creates a fixed-point FIR filter using 16 b to represent the coefficients in fractional format:

```
Hq = qfilt('fir', {b},...'CoefficientFormat', [16,15]);
```

Figure 38 shows the magnitude response of the quantized filter, with the nonquantized magnitude response also shown for comparison. Note that the stopband attenuation for the quantized response is significantly less than 80 dB at various frequency bands. The problem is the poor utilization of the available range for the [16,15] format **(Fig. 39)**.

Two equivalent approaches can be used to make the most of the 16 b. To use the [16,15] format, the coefficients can be scaled by multiplying them by a factor of 8 to make the largest coefficients as close to 1 as possible without overflowing. Alternatively, the [16,18] format can be used so that the quantization range becomes [0.125, 0.125). **Figure 40** shows the magnitude

Continued on page 104

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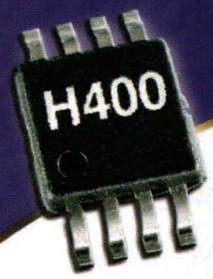

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NEW!	HMC485MS8G	0 LO, High IP3	1.7 - 2.2	0.05 - 0.3	-9	10	+34	$4.49
	HMC410MS8G	High IP3, DBL-BAL	9.0 - 15.0	DC - 2.5	-7.5	40	+24	$4.55
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RCS Measurements Detect Power Lines

Radar-cross-section (RCS) measurements at microwave and millimeter-wave frequencies can be useful in detecting power-line backscattering from low-flying aircraft.

Power lines can be hazardous to low-flying aircraft. Fortunately, it may be possible to use microwave and millimeter-wave radar systems to accurately detect power lines, and prevent accidents. Even under conditions of fog, rain, and snow, it should be possible to measure the polarimetric radar backscattering from a power line using microwave or millimeter-wave radar systems. A main concern, addressed in this study, is the effect of thick ice on these measurements.

Studies have been performed on the use of electro-optical laser radar for collision warning.[1] Unfortunately, the laser radar has numerous shortcomings, including limited range, significant atmospheric attenuation under inclement weather conditions, and difficulty in automating a detection system to warn a pilot of oncoming power lines which make it impractical for power-line collision-warning systems.

Millimeter-wave radar systems, on the other hand, can be used to detect thin objects like power lines in foggy, cloudy, snowy and rainy conditions. While all microwave and millimeter-wave radars are not suitable for this application, some have been used effectively.[2] In one such case, the millimeter-wave radar used linearly polarized waveforms and modeled transmission lines as long, perfectly conducting cylinders.[3] Since power lines are conducting cylindrical wires, they represent difficult targets for many radars. But because a high-voltage power line is made up of strands of wires in a helical arrangement, backscattering detection can be used to detect the power-line cables.

Due to the radar resolution possible at millimeter-wave frequencies, the helical geometry of the power lines is an important factor in influencing the scattering behavior of the electromagnetic (EM) waves, and this can be used in detecting power-line field emissions in off-specular directions. The surface of the power-line cables is periodic along the axis of the cables and usually the period is only a fraction of the helical pitch.

High-voltage power-line cables are

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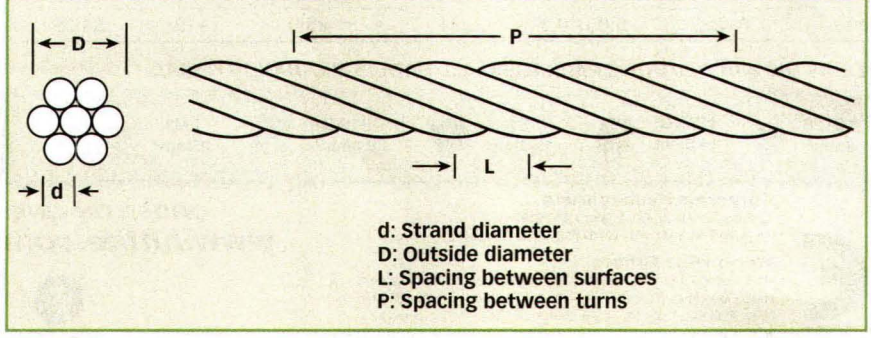

1. In modeling a multistrand high-voltage power-line cable, D is the diameter of the cable, d is diameter of an individual strand, L is the spacing between surfaces, and P is the spacing between turns.

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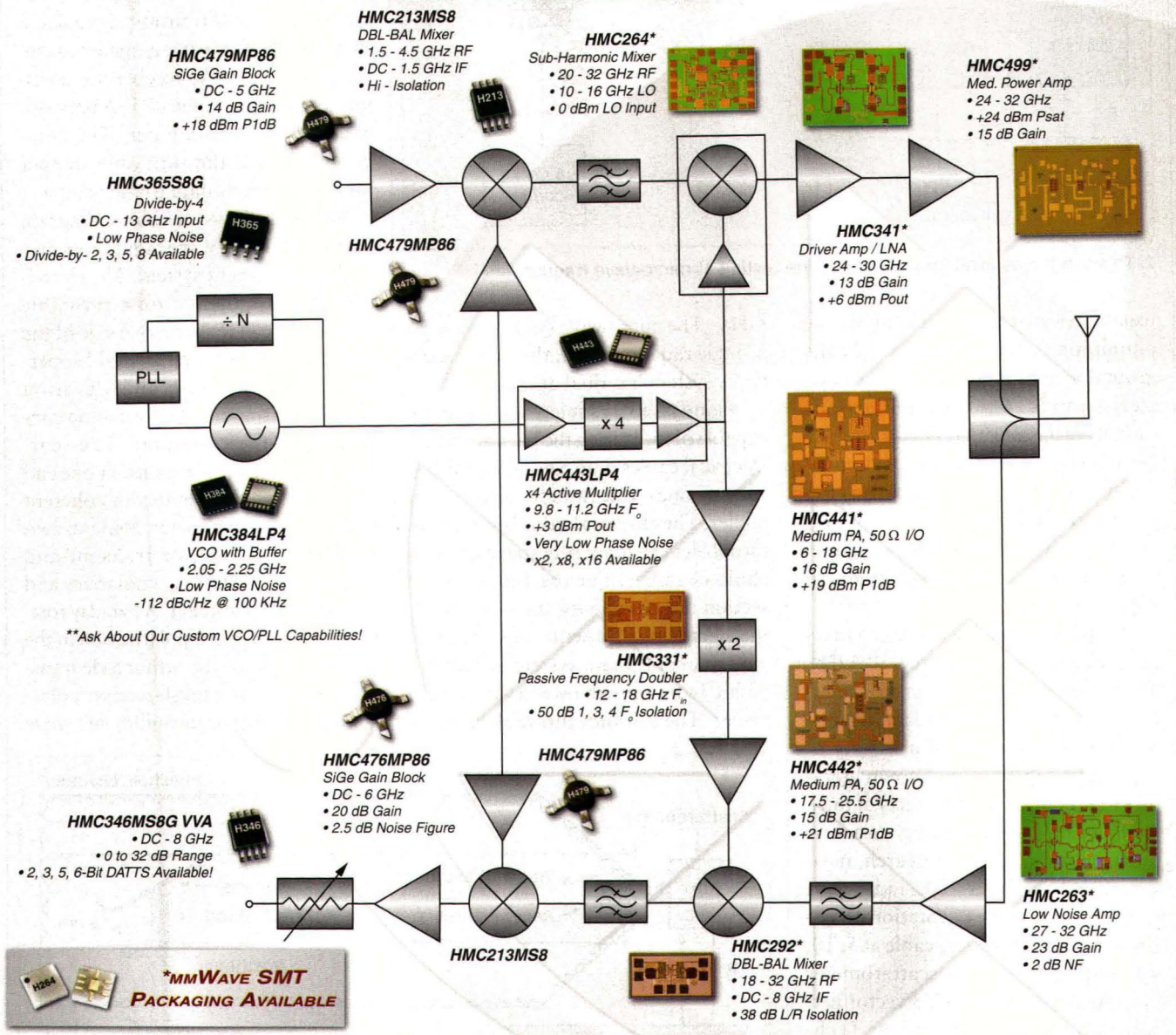

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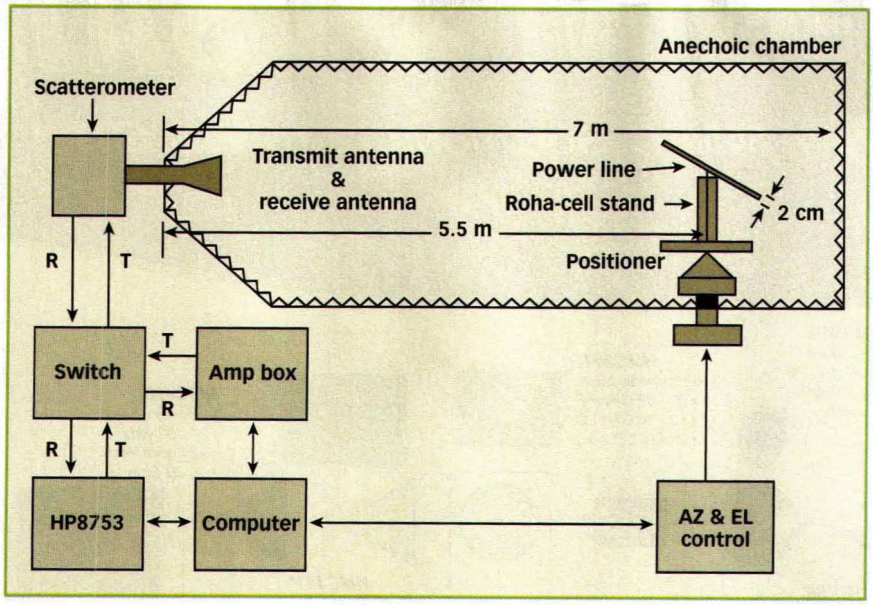


2. This setup was used for RCS power-line testing at microwave frequencies.

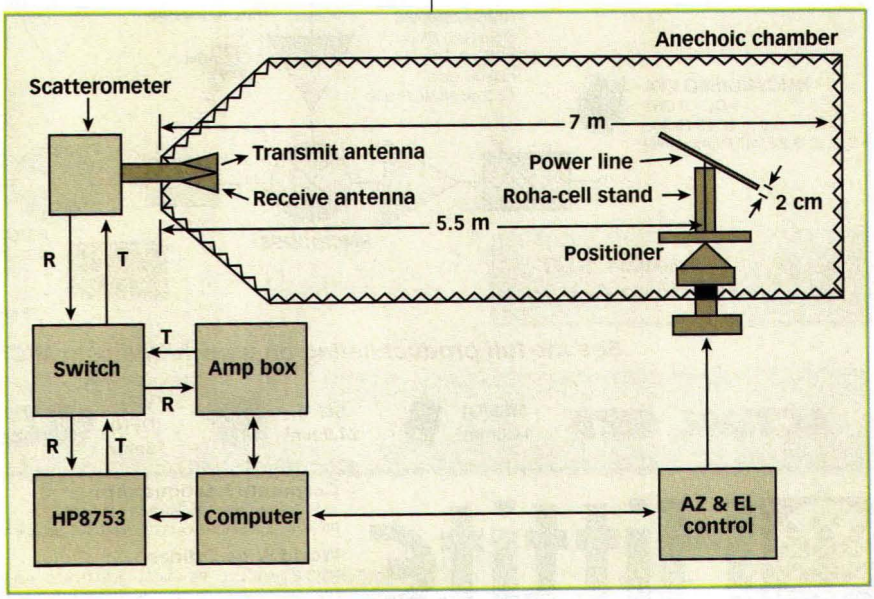
usually constructed from a number of aluminium strands twisted helically around a central core of one or more steel strands. The current-carrying capacity of the cable depends on the number of layers and diameter of aluminium/steel strands. However, in an electrical power distribution network, low-tension and high-current cables are used, made of either aluminium or copper strands.

A typical power-line cable geometry can handle current loads to 420 A **(Fig. 1)**. Although research has been performed on polarimetric radar backscattering measurements of a variety of power line cables,[4-6] no known literature has reported on the effects of thick ice layer on backscattering measurements. For the current research, measurements were made of the radar cross section versus angle of rotation of a 2-cm-diameter power-line cable at 5, 10, 40, and 82 GHz. The scatterometer systems are based on 8753 vector network analyzers from Agilent Technologies (model 8753A for the 5- and 10-GHz bands and model 8753C for the 40- and 82-GHz bands). The measurement systems feature phase and amplitude measurement capabilities and 100-dB dynamic range. The scatterometers used at microwave frequencies (5 and 10 GHz) are slightly different than those used at the millimeter-wave bands (40 and 82 GHz). The microwave system employs a single radar antenna; the millimeter-wave system has dual antennas.

Figure 2 shows a microwave system capable of measuring the backscatter of electrical cables with a good signal-to-background-noise ratio for all incident angles. The electrical cable is mounted on a Roha-cell foam pedestal in an anechoic chamber. For the radar cross-section measurements, the position of the electrical cable with respect to the antenna coordinate system is accomplished by an azimuth-over-elevation positioner. The azimuth turntable is driven by a computer-controlled stepper motor with accuracy of a fraction of a tenth of a degree; the elevation controller is a precise analog positioner.

A transmit/receive antenna for the microwave system consists of an orthomode transducer and a dual-polarized square horn. The scatterometer provides receive and transmit signals to a switch and a microwave amplifier assembly. The microwave receive and transmit signals pass to the 8753A network analyzer and a computer. The computer controls the turntable stepper motor and sends data to the printer.

Figure 3 shows the block diagram of the millimeter-wave radar cross-section measurement system. The electrical cable is mounted on a turntable with Roha-cell foam on one side of the anchoic chamber. The turntable operates in both azimuth and elevation planes with rotational information controlled by the computer. The scatterometer, which is installed at one end of the chamber, operates in coherent mode. The scatterometer, contains two corrugated horns (for transmit and receive), two isolators, a polarizer and a otrthomode transducer. A Faraday rotator is used to achieve polarization in the transmitter while the orthomode transducer is used to establish receiver polarization. Two separate millimeter-wave



3. This measurement setup was used for RCS power-line measurements at millimeter-wave frequencies (40 and 82 GHz).

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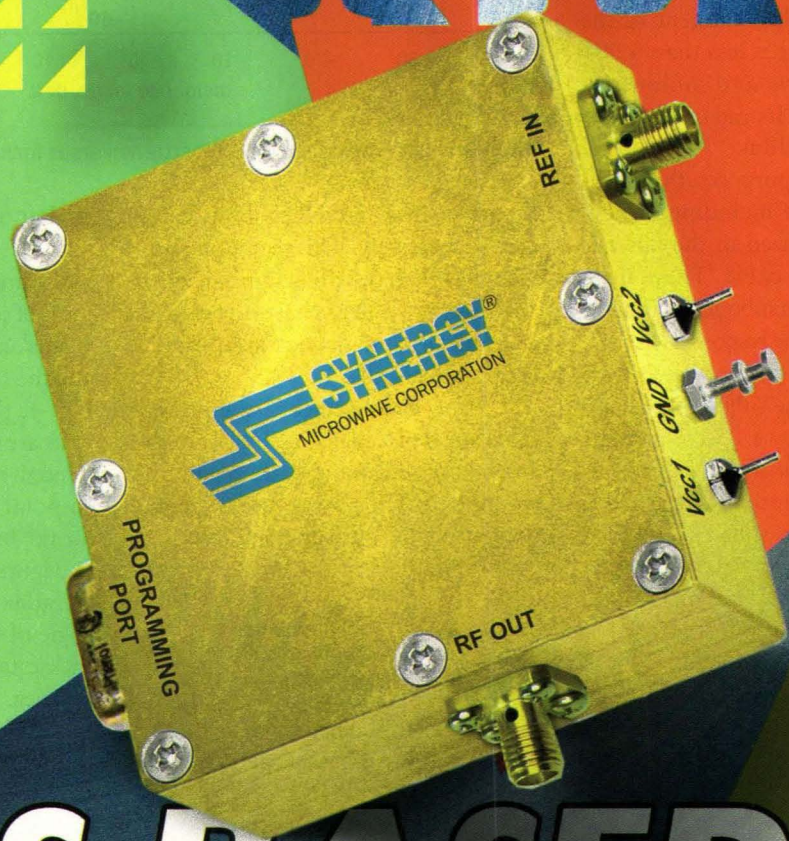

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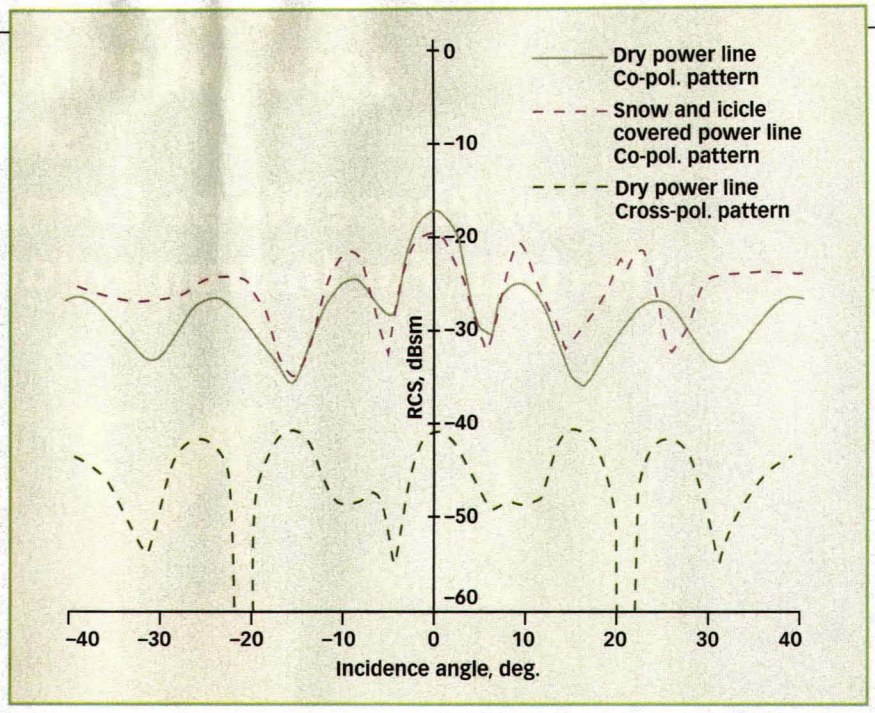
oscillators are employed, operating at 40 and 82 GHz.

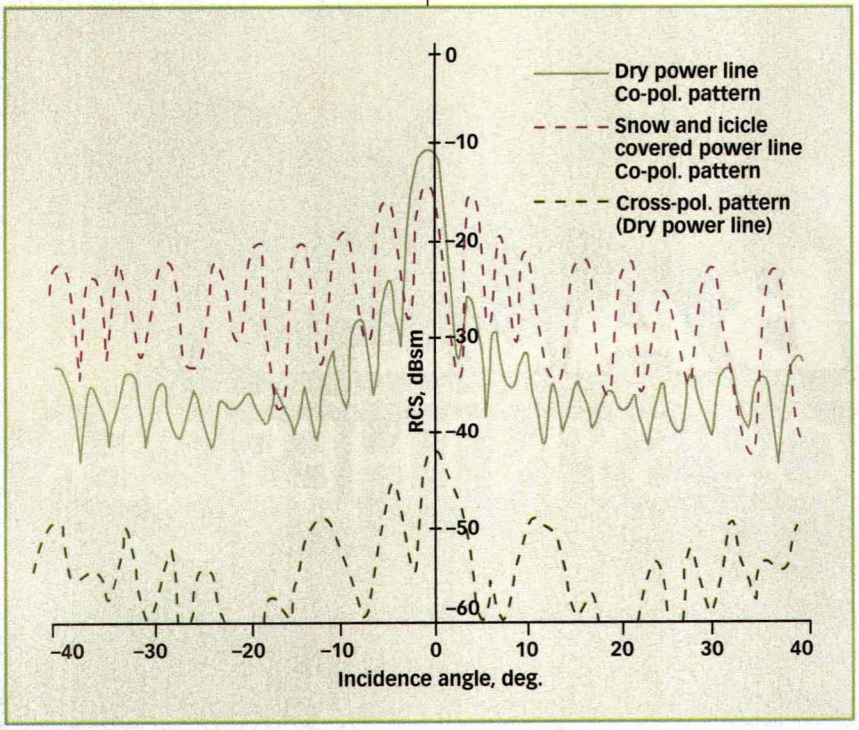
Radar cross-section measurements on the cable were set up in a 7-m-long tapered anechoic chamber; the cable was set a distance of 5.5 m from the receive and transmit antennas. The power-line cable, which was 25 cm long and 2 cm in diameter, was mounted on a Roha-cell foam structure, with the foam structure rigidly attached to the turntable platform. The permittivity of the Roha cell s 1.07 at microwave frequencies and 1.1 at millimeter-wave frequencies. The loss tangent of the Roha-cell structure is less than 0.001 at both microwave and millimeter-wave frequencies. Its radar cross section is below –35 dBm.

For calibration purposes, the chamber and turntable including attachments were measured in the absence of the power-line cable. These measurements were then subtracted from the measured power-line responses. In these measurements, the scatterometers were calibrated using the single-target calibration technique.[7] A 25-cm metallic sphere was used for the microwave 5- and 10-GHz calibration target, while 4- and 2-cm metallic spheres were used for the 40- and 82-GHz calibration targets, respectively. This technique provides less than 0.5 dB magnitude error and less than 5 deg. phase error. In all cases, the signal-to-noise ratio (SNR) was better than 25 dB.



4. These measurements show RCS patterns versus incident angle at 5 GHz.

During measurements, the power-cable line is placed in the H-plane of the antenna system and radiation patterns are plotted. The power-cable orientation during the measurement is similar to that of a radar system mounted on a low-flying aircraft, with power lines in the horizontal plane. **Figure 4** shows the plot of measured radar cross-section versus angle of rotation in the co-polar and the cross-polar planes at 5 GHz. In Fig. 4, the co-polar and cross-polar plots are shown by full and dotted lines, respectively. **Figure 5** shows the measured co-polar and cross-polar at 10 GHz for the braided power line cable (Fig. 1). Figures 4 and 5 show one peak value of radar cross section at an incident angle of 0 deg. There is no significant backscatter at incident angles greater than zero deg. The cross-polar level is less than –35 dB for all angles at 5 and 10 GHz.



5. These measurements show RCS patterns versus incident angle at 10 GHz.

Figure 6 shows measured co-polar and cross-polar radar cross sections versus the angle of incidence at 40 GHz. The co-polar and cross-polar radar cross-section patterns appear in **Fig. 7**. The radar cross-section response of the electric cable at both polarizations (co-polarization and cross-polarization) has peaks at normal incidence and certain discrete incident angles. In a periodic structure such as a power-line cable, Bragg backscatter can be predicted by:

$$\theta_n = \sin^{-1}\left[n\lambda / (2L)\right]$$

where:

λ = the wavelength and

L = the period (shown in Fig. 1).

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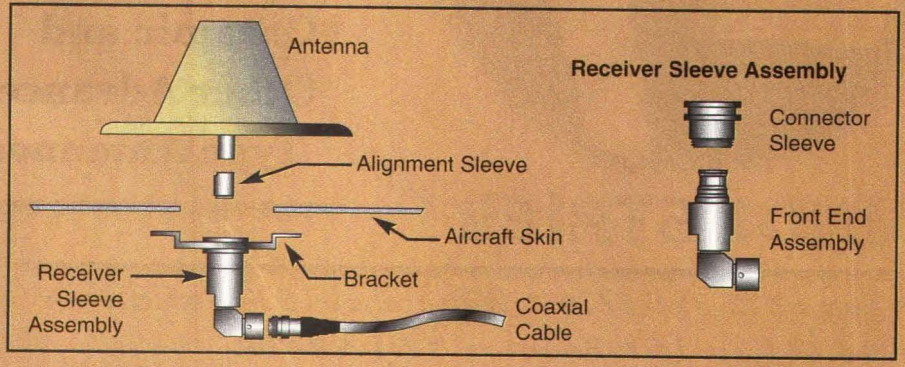

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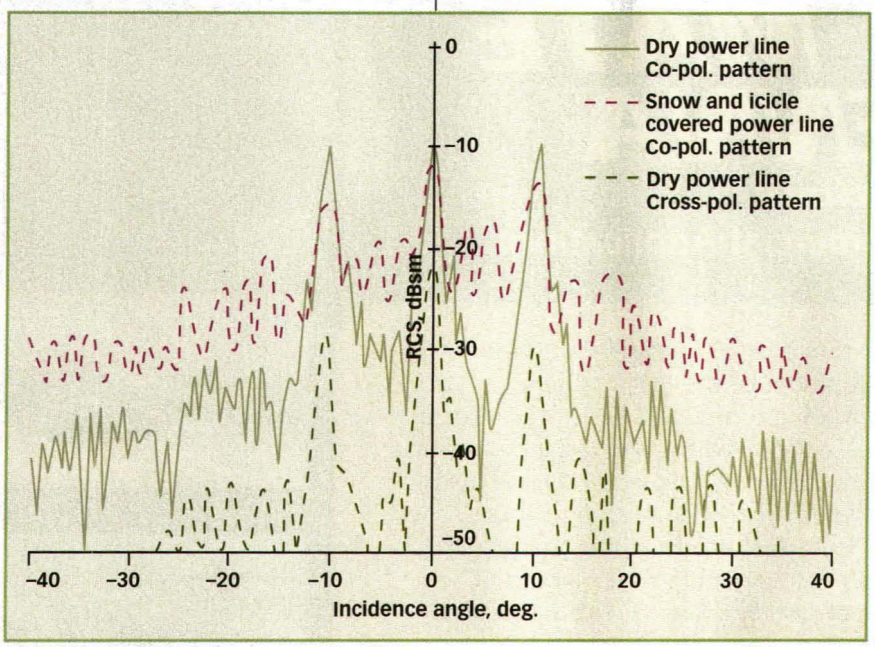
In Figs. 6 and 7, the Bragg scattering ceases at incidence angles larger than 15 deg. Higher-order Bragg modes occur at very low levels at angles away from normal incidence, since most of the scattered energy is in the specular direction.



6. These measurements show RCS patterns versus incident angle at 40 GHz.

In the anechoic chamber, backscattering measurements were performed on a power line covered with a layer of thick ice, at 5, 10, 40, and 82 GHz. A radar transmitter and a power line positioner were set up in the anechoic chamber. During these measurements, the outdoor temperature was about –20°C; the doors and windows of the anechoic chamber were opened during the measurements to maintain the temperature of the chamber at about –50°C. Ice was formed over the power line by spraying it with water and allowing it to freeze. A layer of ice with an average thickness of about 0.5 cm was formed on the horizontal surface of the power-line sample. To simulate real-world ice on outdoor cables, water was dripped onto the braided power-line cable and allowed to freeze. The thick lines in Figs. 4-7 represent the co-polar radar

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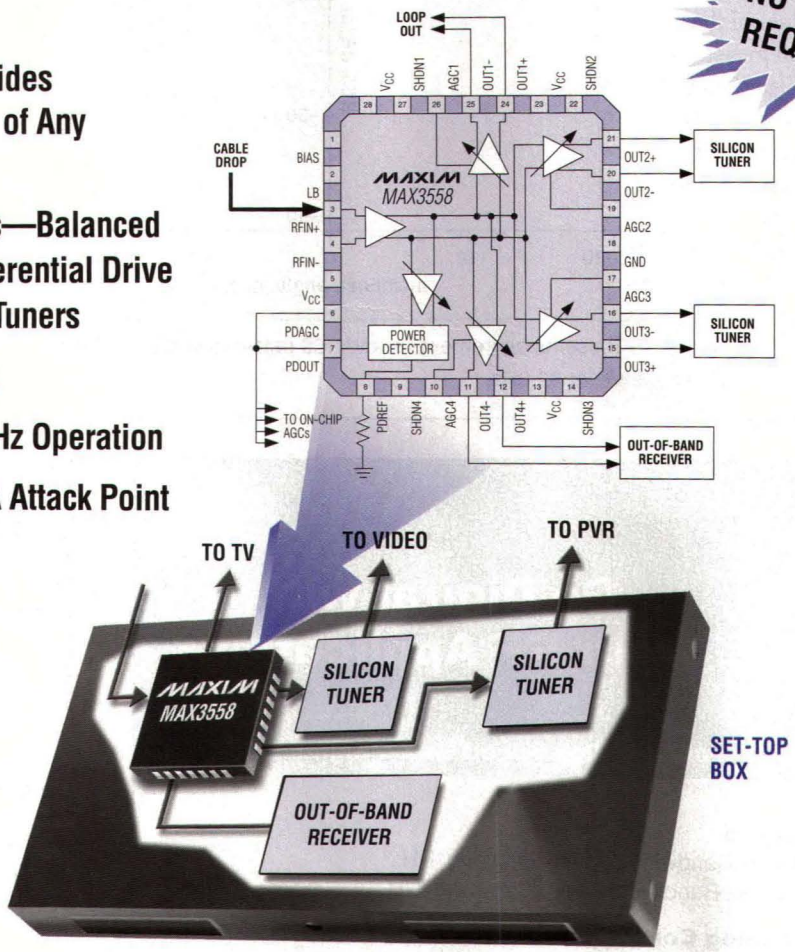

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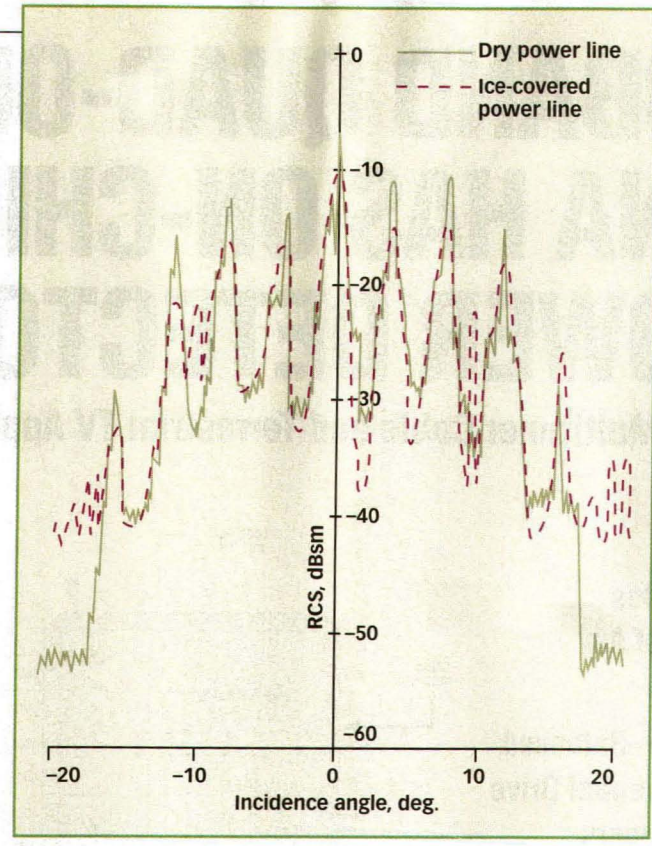
cross-section measurements for the ice-covered power line. The measurements show that ice layers modify the cable-surface reflectivity and the effective surface roughness.

These measurements show that the radar cross section of the ice-covered power line have slightly lower values (a few dB) than the dry power line at angles close to normal incidence, but the values of these radar cross-section measurements increase for higher angles of incidence. These measurements provide a guideline for creating a detection algorithm for power lines using microwave and millimeter-wave radar systems. In conclusion, at microwave frequencies for small diameter power lines ($D/\lambda << 1$), there is no significant backscattering at angles away from normal incidence. MRF



7. These measurements show RCS patterns versus incident angle at 82 GHz.

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For PAs, the performance of E-pHEMT technology offers many well-suited characteristics, including:

1. Saturated drain-source current(I_{dss}) of less than 10 µA at room temperature.
2. Drain current (I_d) of approximately 0 at a gate-source voltage (V_{gs}) of 0.
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4. Superior output power (P_{out}) and high efficiency with bias voltages of less than +3 VDC.
5. No thermal runaway effects (common to bipolar transistors).
6. No secondary breakdown mechanism.
7. The ability to survive under high mismatch conditions.

However, E-pHEMT technology can also provide a combination of high gain, low noise, and wide dynamic range in high-linearity LNA applications, such as intermediate-frequency (IF) amplifiers for commercial communication systems and preamplifiers for magnetic-resonance-imaging (MRI) systems. These types of applications have been made practical with the availability of low-cost plastic-packaged surface-mount E-pHEMT devices specif-

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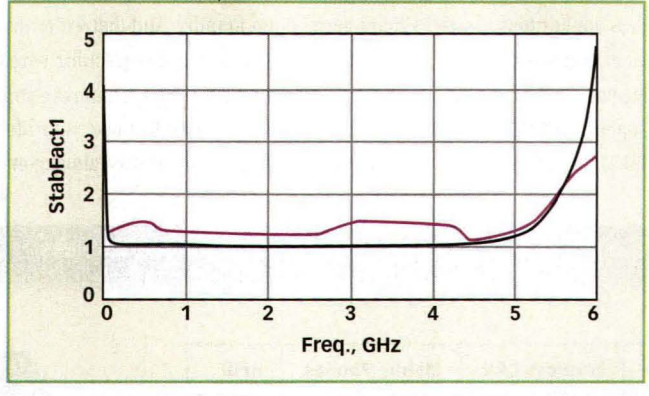

1. The stimulated stability factor (K) of the ATF-54143 E-pHEMT is shown overlaid on the measured K of the device through 6 GHz.

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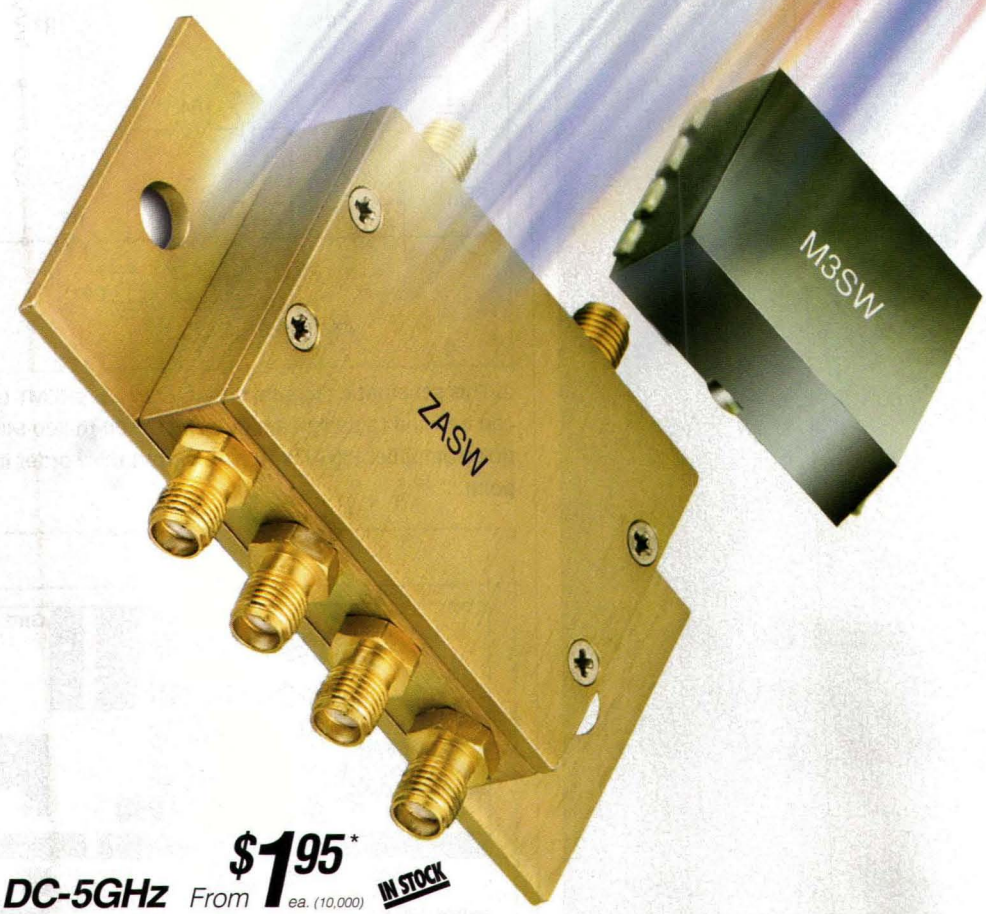

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■ *M3SWA-2-50DR*	*DC-4.5*	*65*	*0.7*	*25*	*4.95 **
					(Qty. 1-9)
• ZASW-2-50DR	DC-5	90	1.7	20	89.95
■ *ZASWA-2-50DR*	*DC-5*	*90*	*1.7*	*20*	*89.95*

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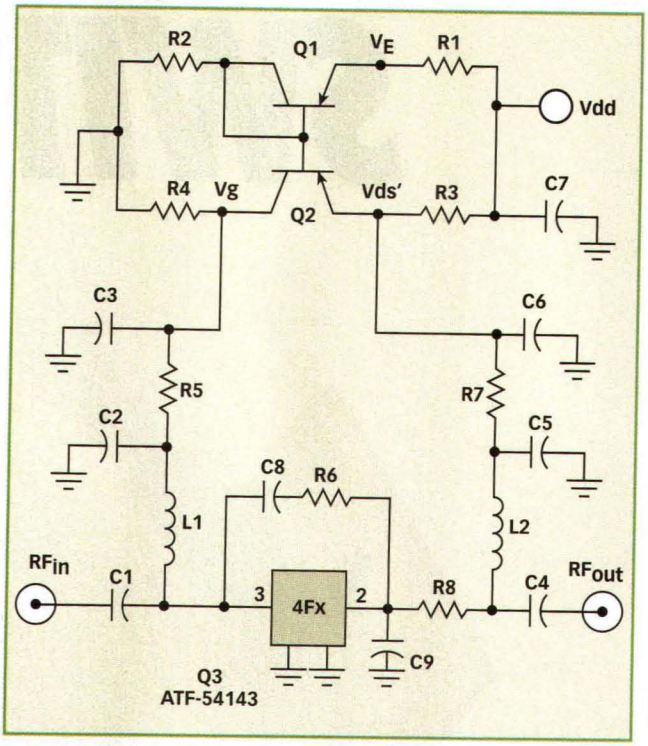
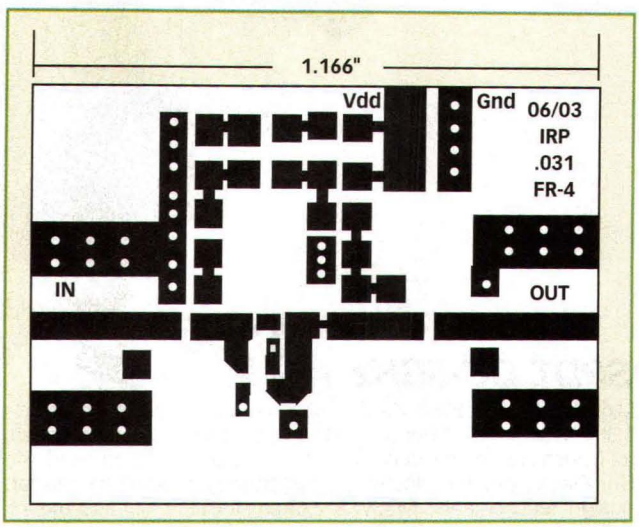


2. This schematic diagram shows how a E-pHEMT transistor can be used to design a multioctave, 100-to-500-MHz low-noise amplifier (LNA) with high output third-order intercept point.



3. This RF layout was used for the demonstration circuit board for the E-pHEMT LNA.

ically designed for LNA applications. This article demonstrates why E-pHEMT technology can economically provide superior electrical performance in the VHF and UHF wireless communications bands commonly associated with other technologies, such as GaAs MESFETs and depletion-mode pHEMTs.

The goal of the design project is to produce a 100-to-500-MHz LNA with an output third-order intercept point (OIP_3) of +36 dBm, a noise figure below 2.0 dB, and gain of 20 dB

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1 GHz	-111	-127	-137	-139	-147
100 MHz	-125	-135	-145	-150	-153

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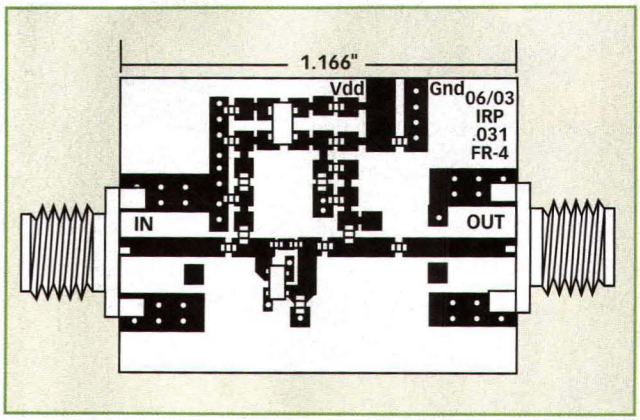

4. This assembly drawing for the 100-to-500-MHz LNA includes input and output connectors as well as ground and bias points.

with flat gain response. Resistive-capacitive (RC) feedback was used to provide good input and output impedance matching to the active device and to ensure unconditional stability. The matching was also required to reduce the overall stage gain to the specified 20 dB level and maintain flat gain across the 400-MHz operating bandwidth. The amplifier design specification includes operation from a +5-VDC supply with current consumption of less than 65 mA.

The ATF-54143 from Agilent Technologies (San Jose, CA) was selected as the active device for the 100-to-500-MHz LNA. The ATF-54143 is one of a family of high-dynamic-range, low-noise enhancement-mode PHEMT discrete transistors designed for use in low-cost commercial applications in the VHF through 6 GHz frequency range. It is housed in a four-lead SC-70 (SOT-343) surface-mount plastic package and operates from a single regulated supply. If an active bias is desirable for repeatability of the bias setting—particularly desirable in high-volume production—the ATF-54143 requires only the addition of a single PNP bipolar junction transistor. Compared to amplifiers using depletion mode devices, the E-pHEMT design has a lower part count and a more compact layout. Besides having a very low typical noise figure (0.5 dB) at 2 GHz, the ATF-54143 is specified at 2 GHz and +3-VDC bias to provides a +36 dBm output third-order intercept point at 60 mA drain current. A data sheet for this device may be downloaded from: http://literature.agilent.com/litweb/pdf/5989-0034EN.pdf

Using the Advanced Design System (ADS) suite of computer-aided-engineering (CAE) software simulation and analysis tools from Agilent/EEsof (Santa Rosa, CA), the amplifier circuit can be simulated in both linear and nonlinear modes of operation. For the linear analysis, transistors can be modeled with a two-port S-parameter file using the *Touchstone* format. More information about Agilent electronic-design-automation (EDA) software may be found at: http://www.agilent.com/eesof-eda .The appropriate ATF54143.s2p file can be downloaded from the *Agilent Wireless Design Center* website:

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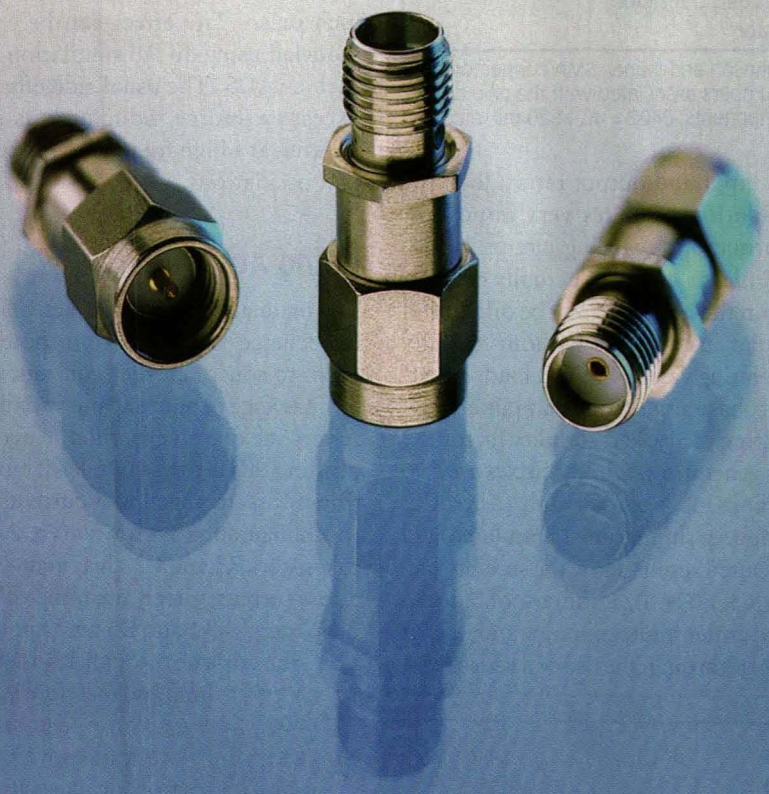

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Table 1: Component lineup for the LNA

COMPONENT	DESCRIPTION	VALUE
C1	0603 chip capacitor	150 pF
C2, C5	0603 chip capacitors	68 pF
C3, C6	0603 chip capacitors	10 nF
C4	0603 chip capacitor	100 pF
C7	0603 chip capacitor	1 μF
C8	0402 chip capacitor	180 pF
C9	0402 chip capacitor	2.2 pF
L1	Toko LL1608-FSR15	150 nH
L2	Toko LL1608-FSR12	120 nH
R1	0603 chip resistor	680 Ω
R2	0603 chip resistor	1300 Ω
R3	0603 chip resistor	22 Ω
R4	0603 chip resistor	270 Ω
R5	0603 chip resistor	47 Ω
R6	0402 chip resistor	680 Ω
R7, R8	0603 chip resistors	4.7 Ω
Q1, Q2	Philips Semiconductor BCV62C transistors	
Q3	Agilent ATF-54143 transistor	

Notes: The input and output RF connectors are EF Johnson end-launch SMA connectors from EF Johnson (part number 142-0701-881). The numbers associated with the chip capacitors and resistors refer to the dimensions of the components: 0402 = 40 × 20 mil, etc.

http://www.semiconductor.agilent.com (type ATF-54143 in the *Quick Search* at the top of the page. Under *Search Results* click on the underlined ATF-54143. Scroll down to the S-parameters listing for 60 mA).

For the nonlinear analysis, a harmonic–balance (HB) simulation was used. The HB simulation was preferred over other nonlinear methods because it is computationally fast, handles both distributed and lumped-element circuitry, and can easily include higher-order harmonics and intermodulation products. The HB approach was used for the simulation of the 1-dB compression point (P_{-1dB}) and OIP_3.

Although this nonlinear transistor model closely predicts the DC and small-signal behavior (including noise), it does not correctly predict the intercept point. To properly model the exceptionally high linearity of the E-pHEMT transistor, a better model was required.

Besides providing information regarding gain, P_{-1dB}, noise figure, and input and output return loss, the simulation provides very important information regarding circuit stability. Unless a circuit is actually oscillating on the bench, it may be difficult to predict instabilities without actually presenting various VSWR loads at various phase angles to the amplifier. Calculating the Rollett stability factor (K) and generating stability circles are two methods made considerably easier with computer simulations. Simulated and measured results show the stability factor, K > 1 **(Fig. 1)**, at the cost of reduced third-order intercept point and output power, through the use of a series resistor on the output.

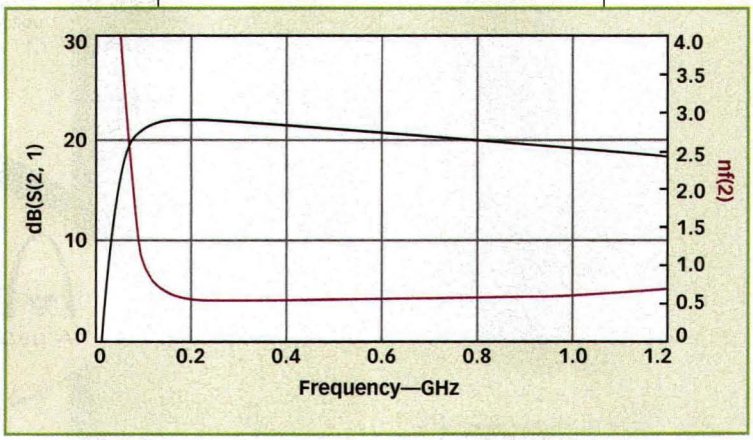
To meet the goals for noise figure, intercept point and gain, the drain source current (I_{ds}) was chosen to be 60 mA. The characterization data in the device data sheet shows that 60 mA gives the best IP_3 combined with a very low minimum noise figure (F_{min}). Also, as shown in the data sheet, a 3-V drain-to-source voltage (V_{ds}) gives a slightly higher gain and easily allows the use of a regulated +5-VDC supply.

The use of a controlled amount of source inductance—usually only a few tenths of a nanohenry—can often be used to enhance LNA performance. This is effectively equivalent to increasing the source leads by approximately 0.025 inch or so. The effect can be easily modeled using an RF simulation tool such as ADS. The usual side effect of excessive source inductance is gain peaking at a high frequency and resultant oscillations.

Using Active Bias

The main advantage of an active biasing scheme is the ability to hold the drain to source current constant over a wide range of temperature variations. A very inexpensive method of accomplishing this is to use two PNP bipolar transistors arranged in a current mirror configuration as shown in **Fig. 2.** Due to resistors R1 and R3, this circuit is not a *true* current mirror, but if the voltage drops across R1 and R3 are kept identical, the current through R3 is stabilized and therefore I_{ds} and V_{ds} are also kept stable. (Note that more information on passive bias networks can be found in Application Note 1222 from Agilent Technologies, at http://literature.agilent.com/litweb/pdf/5988-2336EN.pdf.)

Transistor Q1 is configured with its base and collector tied together. It acts as a simple PN junction, which helps to temperature compensate the



5. The ADS software suite was used to simulate the gain and noise-figure (NF) performance of the E-pHEMT LNA design.

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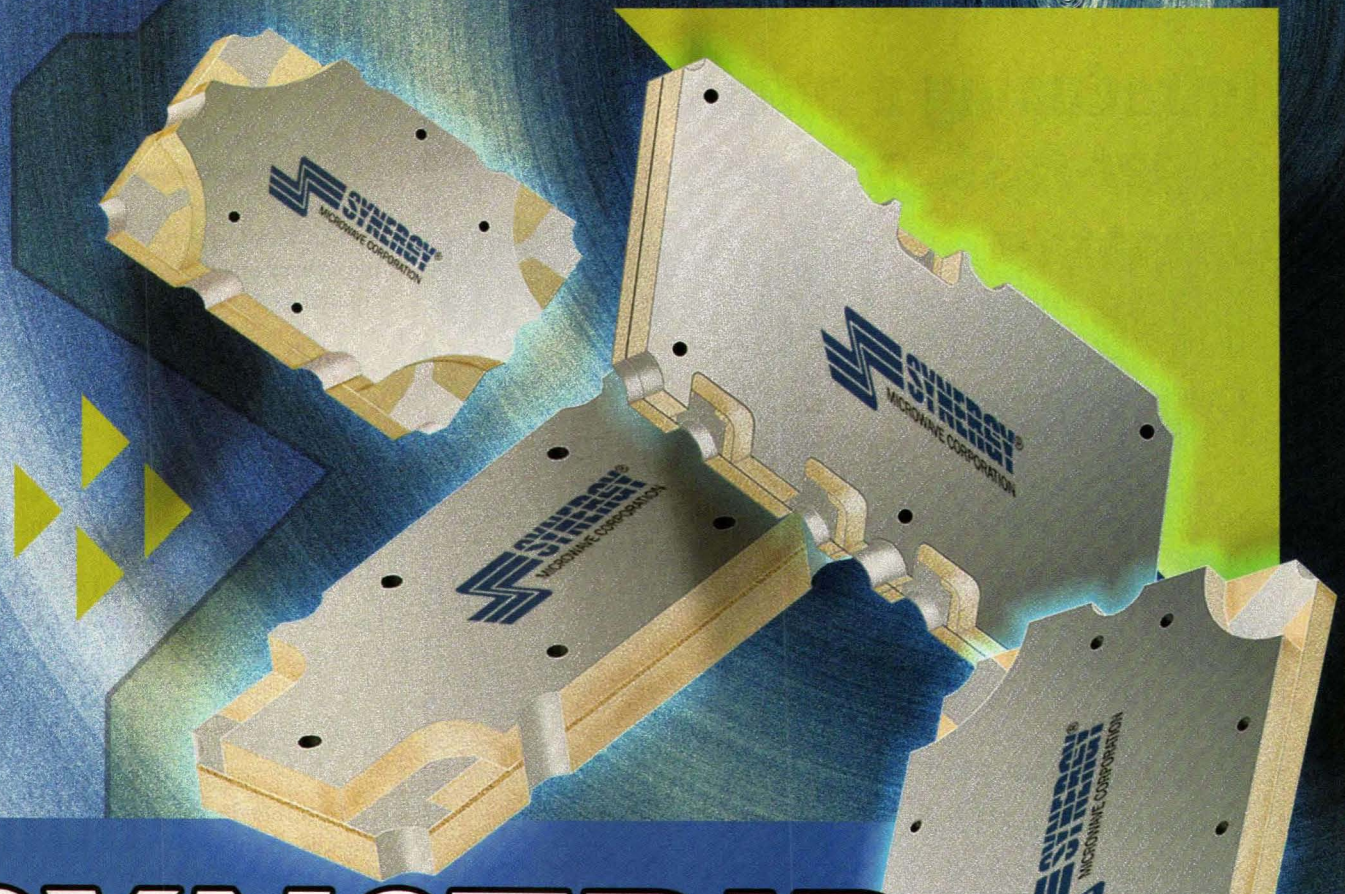

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emitter-base junction of Q2. To calculate the values of R1, R2, R3, and R4 the following parameters must be known or chosen:

I_{ds} = the device drain-to-source current = 60 mA.

I_R = the reference current for active bias = 2.1 mA.

V_{dd} = is the power supply voltage = +5 VDC;

V_{ds} = the device drain-to-source voltage = +3 VDC;

Table 2: Summarizing measurement results

FREQUENCY (MHZ)	GAIN (DB)	NF (DB)	P1DB (DBM)	OIP3 (DBM)
100	20.8	1.20	+16.6	+34.5
200	21.1	0.67	+16.6	+36.3
300	21.4	0.62	+16.6	+36.5
400	21.2	0.61	+16.6	+36.1
500	20.5	0.70	+16.8	+36.5

$V_{ds'}$ = the value of drain-source voltage used in the equations due to the voltage drop across R7 and R8 = +3.56 VDC;

V_{gs} = the typical gate bias = 0.59 V; and

V_{be1} = the typical base-emitter turn-on voltage for transistors Q1 and Q2 = 0.65 V.

Therefore, resistor R3, which sets the desired device drain current, is calculated as follows:

$$R3 = \frac{V_{dd} - V_{ds'}}{I_{ds} + I_{c2}} \qquad (1)$$

where:

I_{C2} is chosen for stability to be 2.1 mA. This value is also equal to the reference current I_R.

The next three equations are used to calculate the rest of the biasing resistors for Fig. 1.

$$R1 = \frac{V_{dd} - V_{ds'}}{I_R} \qquad (2)$$

Note that the voltage drop across R1 must be set equal to voltage drop across R3, but with a current of I_R:

$$R2 = \frac{V_{ds'} - V_{be1}}{I_R} \qquad (3)$$

Resistor R2 sets the bias current through Q1:

$$R4 = \frac{V_g}{I_{C2}} \qquad (4)$$

Resistor R4 sets the gate voltage. $Ic_2 = I_{e2}$ assuming the h_{fe} of the PNP transistors is high. Calculated resistor values differ from actual resistors due to available component values.

Thus, by forcing the emitter voltage (V_E) of transistor Q1 equal to V_{ds},

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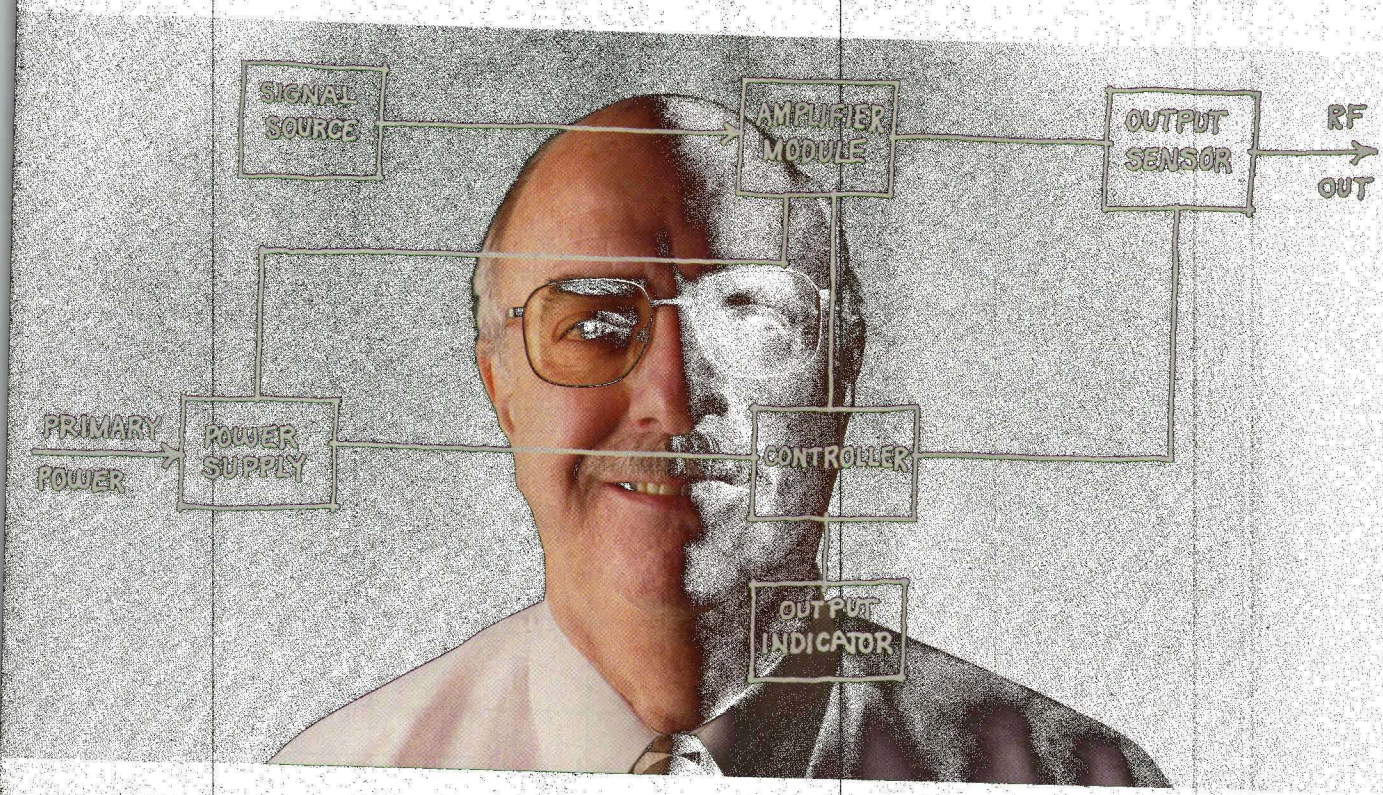

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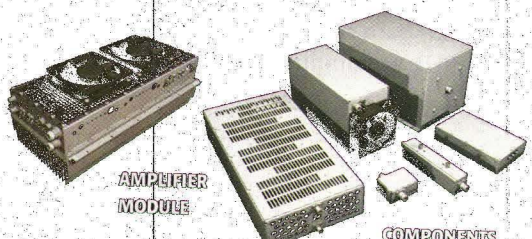

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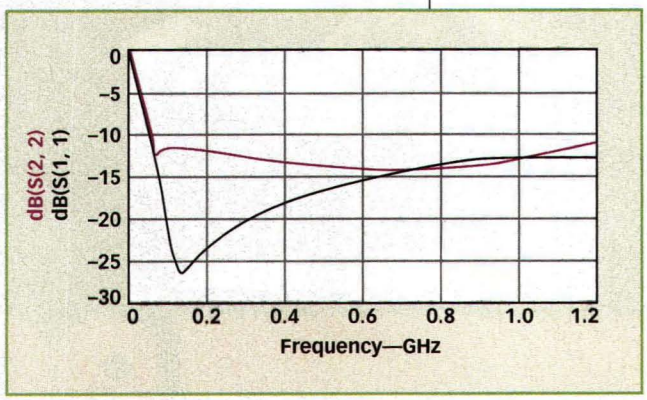
this circuit regulates the drain current in a manner similar to a current mirror. As long as transistor Q2 operates in the forward active mode, this hold true. In other words the collector-base junction of transistor Q2 must be kept reversed biased. **Table 1** shows the parts list of transistors and passive components needed to assemble the LNA.

An evaluation board was designed for the feedback amplifier network. This single-layer board (see **Figs. 3 and 4**) is 0.031-in.-thick FR-4 circuit-board material with a dielectric constant of 4.2. The feedback network should be made as short as possible, since introducing inductance into the feedback network will cause instability in the 5-to-6-GHz region. The RC feedback uses 40 × 20 mil components that are soldered close together with a small solder pad in between.

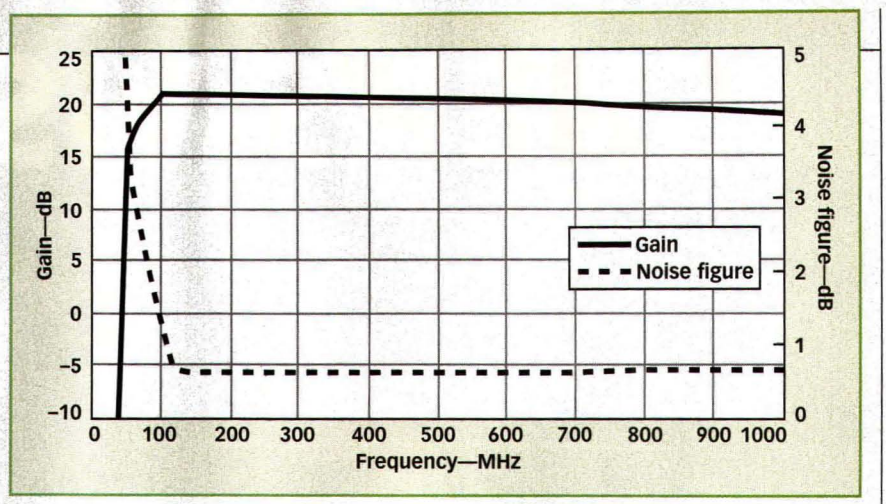
The ATF-54143 is *conditionally* stable below 3.5 GHz, having 29-to-26-dB gain in the 100-to-500-MHz region. The R-C feedback reduces low frequency gain and increases the stability factor to greater than 1 below 2 GHz. The amplifier uses a highpass impedance-matching network, consisting of C1 and L1, for the noise match. The circuit loss will directly relate to noise figure, thus the Q of L1 is extremely important. The LL1608-FSR15 coil from Toko is a small multilayer chip inductor with a rated Q of 19 at 50 MHz. The shunt inductor (L1) provides low frequency gain reduction, which can minimize the amplifier's susceptibility to overload from nearby low-frequency transmitters; it is also part of the input matching network along with C1. Capacitor C1 also doubles as a DC block, while inductor L1 also provides a means of inserting gate voltage for the PHEMT. This requires a good bypass capacitor in the form of C2.

This network represents a compromise between noise figure, input return loss, and gain. Capacitors C2 and C5 provide in-band stability while resistors R5 and R7 provide low-frequency stability by providing a resistive termination. The highpass network on the output consists of a series capacitor C4 and shunt inductors L2, with L2 also providing a means of inserting drain voltage for biasing up the PHEMT. Very short transmission lines between each source lead and ground have been used. The RC-feedback has a dramatic effect on in-band and out-of-band gain, stability, and input and output return loss.

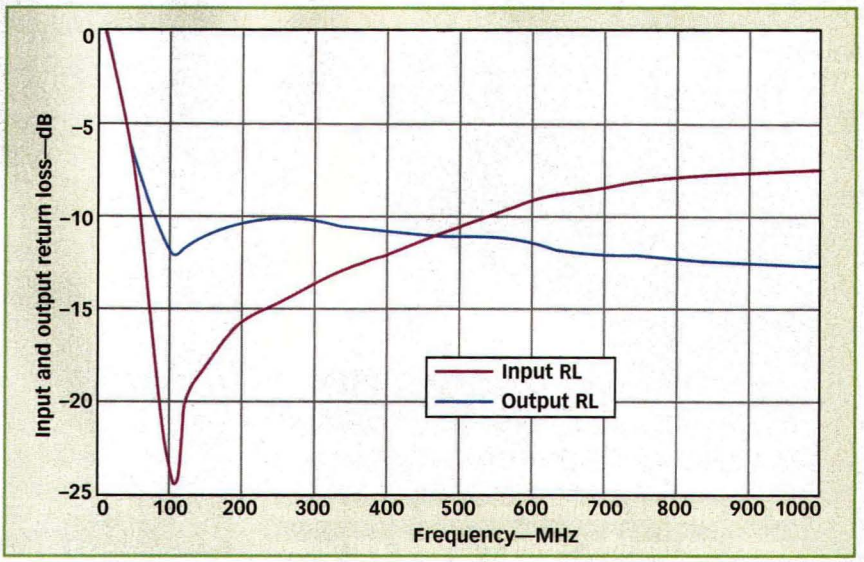
Results from the simulation of gain, NF and for input and output return loss are shown in **Figs. 5 and 6**, respectively. Measured gain and noise figure and input and output return loss appear in **Figs. 7 and 8**, respectively. **Table 2** offers a summary of the measured results for the 100-to-500-MHz LNA. MRF



6. The ADS software suite was also used to predict the input and output return loss for the E-pHEMT LNA.



7. Compare these measured results of gain and noise figure to the simulations of Fig. 5 for the 100-to-500-MHz LNA.



8. Compare these measured results of input and output return loss to the simulations of Fig. 6 for the E-pHEMT LNA.

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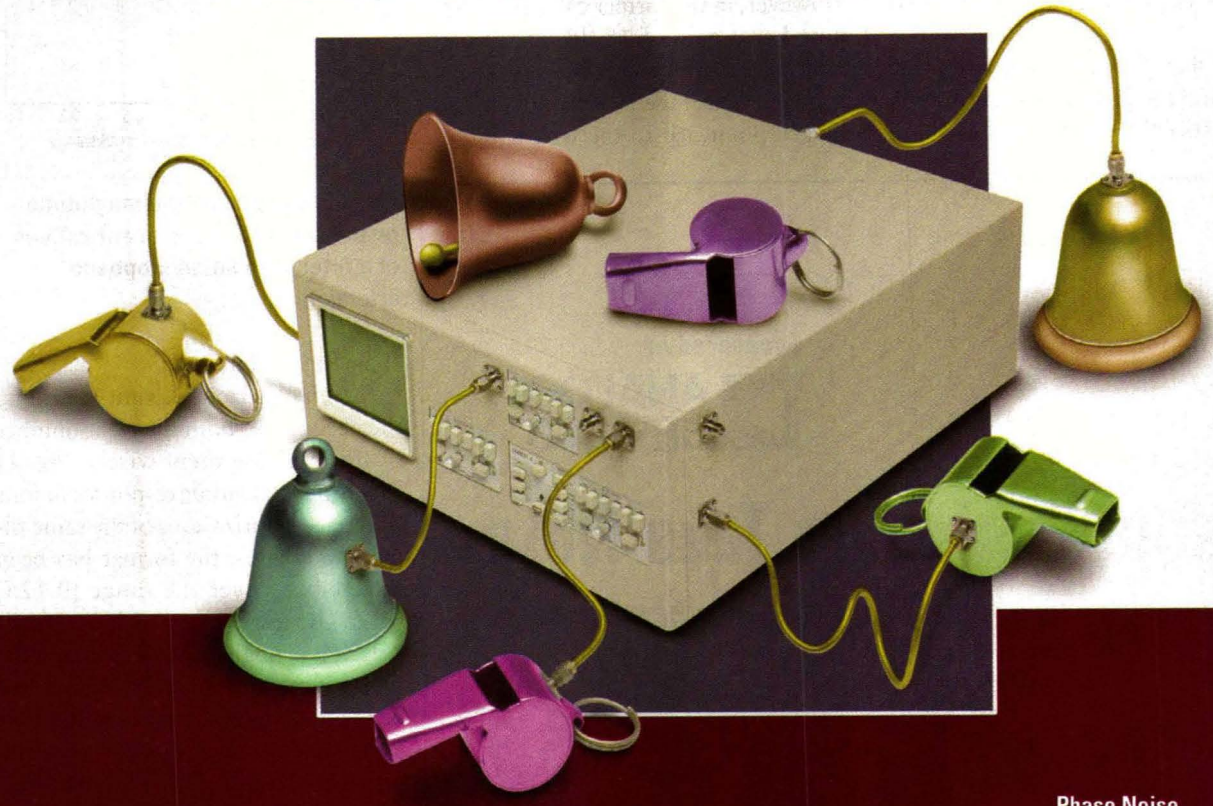

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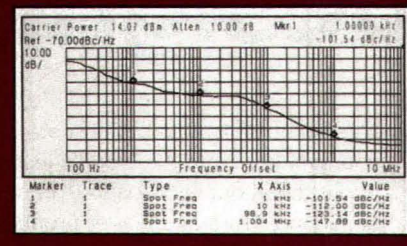

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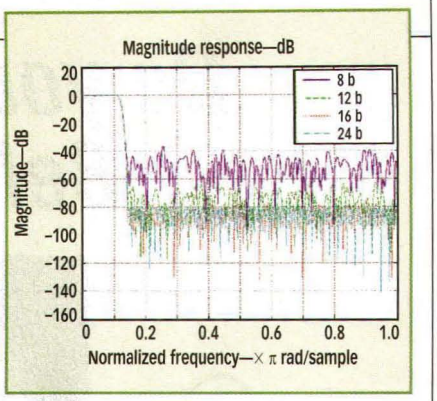
Continued from page 80

response using this format, with an obvious improvement over the first case.

Note that whether the coefficients are scaled and the [16,15] format is used, or the [16,18] format is used without scaling, the actual stored value (the binary bits) of each coefficient is the same. However, in the former case, the filter now has a gain of 18 dB due to the multiplication by eight. But this can be compensated at the end, by moving the binary point 3 b to the left, without changing the bits.



41. This plot shows the magnitude responses for various quantizations of a filter with 80-dB stopband attenuation.

To emphasize the point regarding the need to use both the right number of bits and use them wisely, **Fig. 41** shows the magnitude responses of four different quantizations of the same filter. In all cases, the format has been selected to cover the range [0.125, 0.125]. If fewer than 16 b are available, it is probably wise to redesign the filter, since many more multipliers will be required than be applied with Eq. 6. (If the design specification is changed from 80 dB to 60 dB, 178 multipliers are required as opposed to 220. If it is reduced to 40 dB, 134 multipliers are required. Of course, it is not a given that the application can allow this change in specifications. The point is having less than 16 b makes it unfeasible to attain 80 dB.) On the other hand, increasing the precision to 24 b provides only modest improvements in this case.

In the April 2004 issue, the final installment of this four-part article series on FIR filter design with MATLAB will explore the use of fixed-point arithmetic and how to use a digital-signal-processor (DSP) accumulator to maximum precision in the implementation of digital filters. This final part will also provide a design example based on creating two FIR filters for a GSM digital downconverter. MRF

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14. P.P. Vaidyanathan and T.Q. Nguyen, "Eigenfilters: a new approach to least squares FIR filter design and applications including Nyquist filters," *IEEE Transactions on Circuits and Systems*, Vol. CAS-34, pp. 11-23, 1987.
15. S.K. Mitra, *Digital Signal Processing. A Computer-Based Approach*, 2nd ed., McGraw-Hill, New York, 2001.
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VAT-1	HAT-1	1	1	0.20	0.11	1.10	1.2
VAT-2	HAT-2	2	2	0.20	0.10	1.20	1.2
VAT-3	HAT-3	3	3	0.15	0.12	1.15	1.1
VAT-4	HAT-4	4	4	0.15	0.08	1.15	1.1
VAT-5	HAT-5	5	5	0.10	0.06	1.15	1.1
VAT-6	HAT-6	6	6	0.10	0.02	1.15	1.1
VAT-7	HAT-7	7	7	0.10	0.05	1.15	1.1
VAT-8	HAT-8	8	8	0.10	0.04	1.20	1.1
VAT-9	HAT-9	9	9	0.10	0.02	1.15	1.1
VAT-10	HAT-10	10	10	0.20	0.03	1.20	1.1
VAT-12	HAT-12	12	12	0.10	0.05	1.20	1.1
VAT-15	HAT-15	15	15	0.30	0.05	1.40	1.1
VAT-20	HAT-20	20	20	0.75	0.18	1.20	1.1
VAT-30	HAT-30	30	30	0.30	0.38	1.15	1.1

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Learn More About The Basics Of WLANs

WIRELESS LOCAL-AREA NETWORKS (WLANs) represent one of the fastest-growing segments of wireless technology, expanding from enterprise networks to millions of small/home-office applications. The appeal of the WLAN is obvious, since it allows operators to send up a fully operational network, with computers, printers, and other peripheral devices, without the benefit of wires. For those who need a basic but informative tutorial introduction to WLANs, Atmel Corp. (San Jose, CA) offers a free, eight-page application note, "A Tutorial on WLANs" on their website at www.atmel.com.

The application note details the many advantages of WLANs over traditional wired Ethernet-type networks, including greater mobility, the speed and simplicity of system integration, portability (the ease of modifying or moving an entire network system when a company moves), scalability, and reduced cost compared to wired networks.

The note includes a review of WLAN operational basics, presenting a graphical image of a typical WLAN physical configuration, and compares the different types of technical, including RF, microwave, and infrared signals, that are used to implement a WLAN system. The note concisely compares the differences between direct-sequence-spread-spectrum (DSSS) and frequency-hopping-spread-spectrum (FHSS) approaches to a WLAN system, and how each approach fares with interference and noise as well as with in-building multipath fading effects. The note also covers such aspects as interference from other emitters, security provisions, variations in the propagation path that can alter the effective range of a WLAN system, the data throughput that can be expected for a given WLAN approach, the safety (what kind or health issues are expected from the RF power transmitted in a WLAN system), and the cost of operating a WLAN system.

It should be noted that this is a very basic introduction and overview of WLAN systems, but for those in need of starting point, this provides an easy-to-follow, non-mathematical explanation. Download a free copy from the company's website.

Atmel Corp., 2325 Orchard Pkwy., San Jose, CA 95131; (408) 441-0311, FAX: (408) 487-2600, Internet: www.atmel.com.

The short application note provides a quick overview of the handheld spectrum analyzer as well as the proper use of the probes for magnetic- and electric-field measurements.

Simplifying EMC Precompliance Testing

ELECTROMAGNETIC COMPATIBILITY (EMC) is a major issue in most countries during the development of an electronic product. A number of regulatory agencies around the world, including the Federal Communications Commission (FCC) in the United States and CISPR for members of the European Union, set standards for acceptable amounts of radiated emissions by product categories. Before an electronic product can be finalized for public sale, it must pass compliance testing in an accredited test laboratory or facility. However, since securing time in such facilities is expensive, many designers prefer to perform precompliance EMC testing during the development stages of a product, in order to understand both the effects of emissions on the product under development and the sources of emissions from the product itself. To aid in the EMC precompliance process, Bantam Instruments (Sunnyvale, CA) offers a handy two-page applications note (note 401B-1) "EMC Pre-Compliance Made Easy."

The two-page note is based on the company's extremely portable model 401B personal spectrum analyzer, a handheld unit that works with the company's E- and H-field probes to make near- and far-field measurements from 30 to 1024 MHz. The E-field sensitivity of the probe/analyzer combination is about 57 dBμV/m across the full band while the H-field sensitivity of the probe/instrument combination is about 57 dBμV/m at 1024 MHz decreasing to about 87 dBμV/m at 30 MHz.

The short application note provides a quick overview of the handheld spectrum analyzer as well as the proper use of the probes for magnetic (H) field and electric (E) field measurements. In addition, it reviews the use of an active antenna (with complementary frequency range of 30 to 1024 MHz) and a preamplifier intended for use with the EMC antenna.

Copies of the two-page application note are free upon request from:

Bantam Instruments, 197 South Murphy Ave., Sunnyvale, CA 94086; (408) 736-3030, FAX: (408) 904-5221, Internet: www.bantaminstruments.com.

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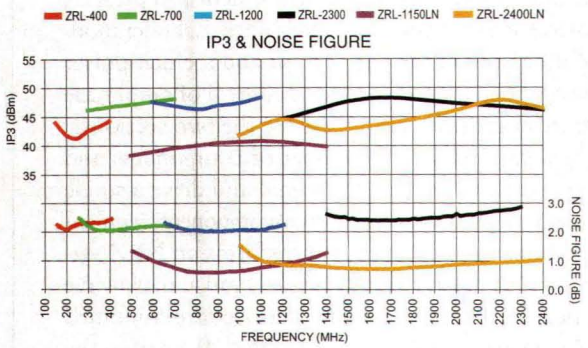

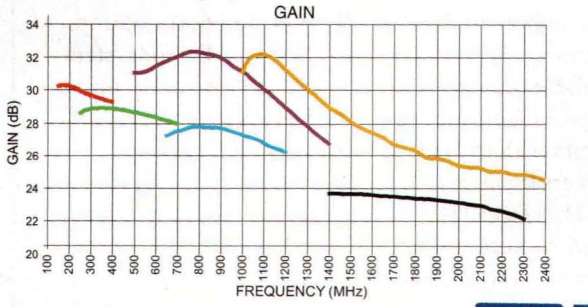

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ZRL-1200	650-1200	27	2.0	46	24.3	119.95
ZRL-2300	1400-2300	24	2.5	46	24.6	119.95
ZRL-2400LN	1000-2400	27	1.0	45	24.0	139.95

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cover story

Module Merges 11.5-Gb/s NRZ/RZ Converter, Driver

Integration of the data converter and high-speed driver spares optical system integrators the task of optimizing performance with dissimilar components.

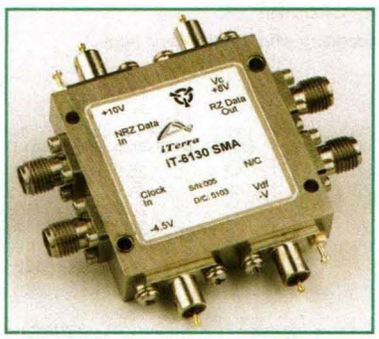
Long-haul optical communications systems rely on a variety of electrical and optical components to send and receive high-rate data. Among these components, data converters and modulators represent important functions that are also technically difficult to integrate. Fortunately, iTerra Communications (Palo Alto, CA) has developed its model iT6130 (see table), a single surface-mount module that combines a non-return-to-zero (NRZ)/return-to-zero (RZ) converter and a modulator driver for long-haul applications to 11.5 Gb/s.

PAOLA TABACCO
Senior Engineer
iTerra Communications, 2440-A Embarcadero Way, Palo Alto, CA 94303; (650) 424-1937, FAX: (650) 424-1938, e-mail: Paolo.tabacco@iterrac.com, Internet: www.iterrac.com.

Modulation techniques that extend the distance between optical amplifiers can reduce hardware costs in long-haul optical communications systems. RZ modulation, the leading technique for long-haul optical transmissions, has traditionally been difficult to implement in practical systems. At present, optical RZ pulses can be generated by separate cascaded Mach-Zehnder modulators driven by an NRZ data stream for one section and a clock pulser for the second section. The approach requires precise control of amplitude and phase, as well as separate microwave amplifiers for the two sections.

A simpler approach which can reduce the number of components and optical loss would be to generate an electrical RZ signal and drive a single Mach-Zehnder optical modulator. Two of the fundamental components required for this technique are a converter that electrically transforms an NRZ signal, typically coming from a multiplexer, to an RZ signal, and a microwave amplifier that drives an optical modulator. These two devices are combined in the iT6130 NRZ-to-RZ converter/modulator driver **(Fig. 1)**, sparing the optical system integrator component matching, tuning, and optimization.

Combining these two functions produces significant benefits for system integrators. Optical modulator drivers are notoriously sensitive to the input from the amplifier that drives them, and matching the two is extremely difficult if the signal is RZ rather than NRZ because of its different spectral shape. The modulator drivers themselves are equally sensitive to the signal conditions at their inputs. The iT6130 eliminates the device selection, design, and optimization required to accommodate these demands, because all required functions are contained within the module and optimized to work with each other. The result is a data converter/driver amplifier that is smaller and lighter than a unit built from discrete parts and typically delivers bet-



1. The iT6130 module combines an NRZ-to-RZ converter and a high-speed modulator driver for long-haul applications to 11.5 Gb/s.

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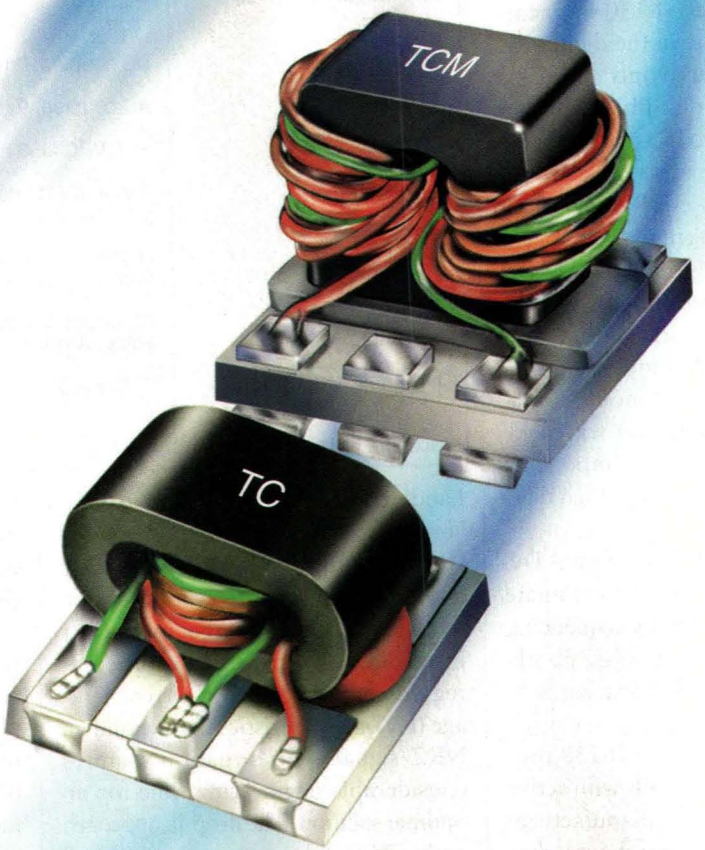

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TC1-15	1C	800-1500	800-1500	1.29
TC1.5-1	1.5D	.5-2200	2-1100	1.59
TC1-1-13M	1G	4.5-3000	4.5-1000	.99
TC2-1T	2A	3-300	3-300	1.29
TC3-1T	3A	5-300	5-300	1.29
TC4-1T	4A	.5-300	1.5-100	1.19
TC4-1W	4A	3-800	10-100	1.19
TC4-14	4A	200-1400	800-1100	1.29
TC8-1	8A	2-500	10-100	1.19
TC9-1	9A	2-200	5-40	1.29
TC16-1T	16A	20-300	50-150	1.59
TC4-11	50/12.5D	2-1100	5-700	1.59
TC9-1-75	75/8D	0.3-475	0.9-370	1.59

LEADS Plastic Base

(actual size) MODEL	Ω Ratio & Config.	Freq. (MHz)	Ins. Loss❖ 1dB (MHz)	Price $ea. (qty. 100)
TCM1-1	1C	1.5-500	5-350	.99
TCML1-11	1G	600-1100	700-1000	1.09
TCML1-19	1G	800-1900	900-1400	1.09
TCM2-1T	2A	3-300	3-300	1.09
TCM3-1T	3A	2-500	5-300	1.09
TTCM4-4	4B	0.5-400	5-100	1.29
TCM4-1W	4A	3-800	10-100	.99
TCM4-6T	4A	1.5-600	3-350	1.19
TCM4-14	4A	200-1400	800-1000	1.09
TCM4-19	4H	10-1900	30-700	1.09
TCM4-25	4H	500-2500	750-1200	1.09
TCM8-1	8A	2-500	10-100	.99
TCM9-1	9A	2-280	5-100	1.19

Dimensions (LxW): TC .15" x .15" TCM .15" x .16" ❖Referenced to midband loss.

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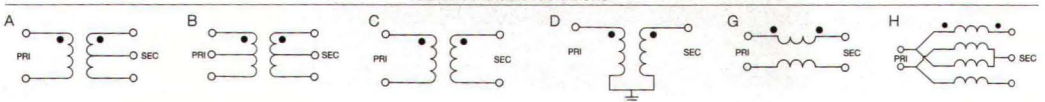

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MODULE

ter performance. The system integrator need be concerned only with the digital interface at the iT6130's input and the output interface. The latter interface is simplified in the iT6130 by incorporating a DFF core in the digital RZ converter. A user need only adjust the NRZ data and clock phase margin which can be as broad as 320 deg.

The iT6130 is a complete subassembly for generating a high-level RZ signal from clock and NRZ data inputs. Its output is adjustable from 4 to 7 VDC p-p and tailored for use with lithium-niobate (LiNo) modulators. Maximum data rate is 11.5 Gb/s. The module consists of a single-chip GaAs monolithic converter and a hybrid traveling-wave FET amplifier plus ancillary components mounted on a substrate with high thermal conductivity.

The subassembly has internal DC regulators and filtering that eliminate the need for power-supply sequencing and accommodate even noisy, poorly regulated DC inputs. If interfacing agility or timing-dependent jitter optimization is required, the iT6130 provides the desired phase shift with active phase shifters, the only manufacturer yet to do so. This approach was chosen because cascaded passive phase shifters produce loss that must be compensated for with amplification, which increases cost and complexity.

RZ modulation enables link extension because of its inherently higher peak power for a given input signal. If the average optical power is constant, RZ modulation will produce twice the peak power of NRZ modulation (but require twice the bandwidth). Consequently, RZ modulation produces a theoretical 3-dB advantage in receiver sensitivity (reduced in practice by 0.5 dB because of increased shot noise caused by greater peak power in the receiver). While other factors complicate this rule, an RZ-coded signal can still allow much longer span lengths than NRZ.

However, implementing RZ modulation creates engineering problems. All components, from RZ converter to receive filter, must handle twice the bandwidth of an NRZ signal at the same data rate. "Off-the-shelf" RZ driver amplifiers are rare and expensive because they must accommodate a driving signal that forces their input stage to operate in a different region of their transistor current-voltage (I-V) characteristics than would an NRZ signal. This situation requires considerable engineering time for an optimal solution. The drop-in approach of the iT6130 should be welcome by all designers who have themselves wrestled with optimization and integration.

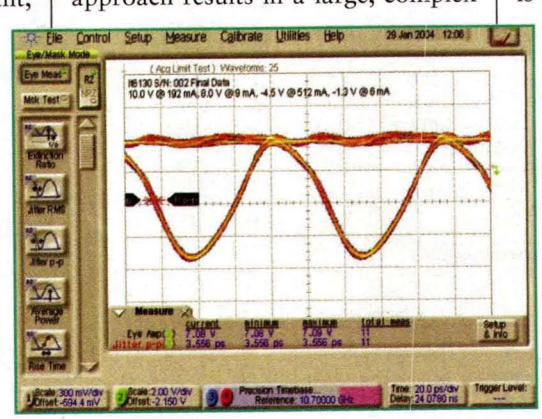
RZ signals were initially achieved with optical methods, in which electro-optical modulators were cascaded, and the pulse was generated in the optical domain using an optical filter. This approach results in a large, complex device that is expensive to produce, draws considerable current, and generates loss in the second optical modulator that requires a higher-power laser for compensation, but achieves very low jitter. Manufacturers have worked to enhance electrical RZ generation in recent years, since the method requires only a single optical modulator, modulator driver, and bias controller, to reduce system costs by $3000 to $5000. Only one driver amplifier is required for reduced power dissipation.

With "electrical RZ," the NRZ data from the multiplexer is converted to an RZ signal, amplified, and sent to the optical modulator as a 5-to-7-VDC signal. The output signal must have a flat zero-level (rail) and low jitter, which is a difficult challenge. Jitter performance of the RZ modulator integrated within the iT6130 represents some of the best jitter characteristics yet achieved, producing eye diagrams that are virtually indistinguishable from those produced by optical techniques **(Fig. 2)**. Jitter at an output of 7 VDC is 3.5 ps p-p.

Two versions of the iT6130 are available: an SMA module and a two-chip surface-mount version. **iTerra Communications, 2440-A Embarcadero Way, Palo Alto, CA 94303; (650) 424-1937, FAX: (650) 424-1938, e-mail: Mahvish_bari@iterrac.com, Internet: www.iterrac.com.**

The iT6130's specifications at a glance

Function	Converts NRZ to RZ data signal
Maximum data rate	11.5 Gb/s
RZ data output level	4 to 7.5 VDC p-p
Input amplitude (max)	+23 dBm
Inputs	
NRZ data	DC coupled, 9 to 11 Gb/s, 0 to –0.8 VDC
Clock	AC coupled, 9 to 11 GHz synchronous with NRZ input
Power	–4.5 VDC (550 mA), +10 VDC (200 mA)
Control	5 to 8 VDC to control output level from 4 V p-p to 6.7 V p-p (<20 mA)
Jitter control	–0.8 to –3 VDC to adjust output duty cycle (pulse width) (<10 mA)
Outputs	
RZ data	AC coupled. Inverted RZ where 0 is high, 1 is negative-going. Amplitude adjustable from 4 to 6.7 VDC p-p.
Operating temperature	0° to 85°C
Dimensions	
Connectorized	1.5 ×1.5 × 0.375 in.



2. The jitter performance of the iT6130 (3.5 ps p-p at 7 VDC) is nearly identical to performance achieved by an optically generated RZ signal.

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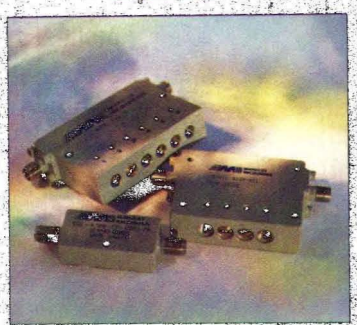

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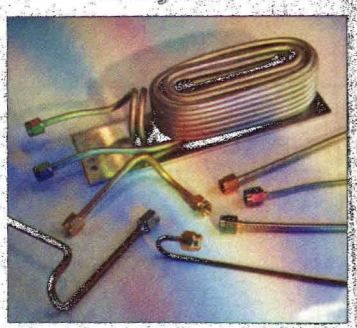

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CMOS Power Amp Drives Dual GSM Bands

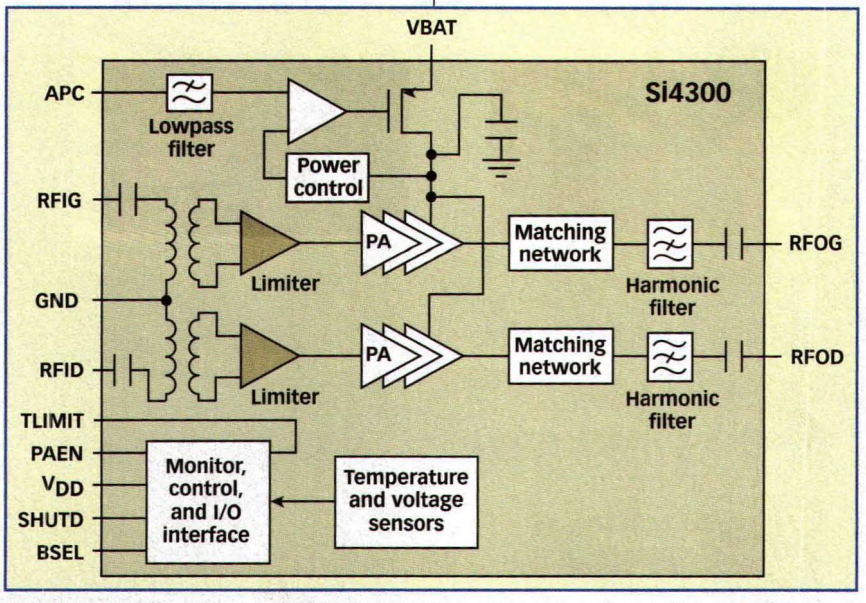
A clever distributed circuit design helps overcome the power limitations of conventional silicon CMOS processing in providing high GSM handset power levels with generous efficiency.

JACK BROWNE
Publisher/Editor

Amplifier specifiers involved with selecting products for GSM cellular-telephone designs have learned to choose from a substantial list of module solutions. These modules are accompanied by one or two amplifier chips and a power-control IC, and sometimes still require external circuitry for DC blocking and impedance matching. In contrast, the Si4300 GSM/GPRS power amplifier (PA) from fabless-semiconductor company Silicon Laboratories (Austin, TX) is a complete monolithic amplifier integrated circuit (IC), a single device that includes all the functions between a GSM handset transceiver and the antenna/switch module. Operating in both the GSM 900 (880-to-915-MHz) and DCS 1800 (1710-to-1785-MHz) bands, the IC includes the amplifiers, harmonic filtering, complete power-control circuitry, and impedance matching and requires no external discrete components.

Unlike most (GaAs) GSM handset PAs, the Si4300 **(see figure)** is made with a standard 0.35-µm silicon CMOS process. By using a novel circuit design employing distributed device stages, the Si4300 can achieve the high voltages needed for high-gain amplification in CMOS. Unlike GaAs-based PA modules which can measure 7 × 10 mm or more, the Si4300 is supplied as a single die in a compact 3.9 × 6.4 mm ceramic package. Like the GaAs PA modules, the Si4300 is designed for voltages from +3.0 to +5.5 VDC.

The Si4300 features a GSM output-power level of +34.7 dBm and DCS output-power level of +32.3 dBm, while achieving typical power-added efficiency of better than 50+ percent. **Silicon Laboratories, Inc., 4635 Boston Lane, Austin, TX 78735; (512) 416-8500, (877) 444-3032, FAX: (512) 464-9444, e-mail: PAinfo@silabs.com, Internet: www.silabs.com.**



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Gali 21	DC-8000	14.3	12.6	4.0	27	128	40	3.5	.99
Gali 2	DC-8000	16.2	12.9	4.6	27	101	40	3.5	.99
Gali 33	DC-4000	19.3	13.4	3.9	28	110	40	4.3	.99
Gali S66	DC-3000	22	2.8	2.7	18	136	16	3.5	.99
Gali 3	DC-3000	22.4	12.5	3.5	25	127	35	3.3	.99
Gali 6F	DC-4000	12.1	15.8	4.5	35.5	93	50	4.8	1.29
Gali 4F	DC-4000	14.3	15.3	4.0	32	93	50	4.4	1.29
Gali 51F	DC-4000	18.0	15.9	3.5	32	78	50	4.4	1.29
Gali 5F	DC-4000	20.4	15.7	3.5	31.5	103	50	4.3	1.29
Gali 55	DC-4000	21.9	15.0	3.3	28.5	100	50	4.3	1.29
Gali 52	DC-2000	22.9	15.5	2.7	32	85	50	4.4	1.29
Gali 6	DC-4000	12.2	18.2	4.5	35.5	93	70	5.0	1.49
Gali 4	DC-4000	14.4	17.5	4.0	34	93	65	4.6	1.49
Gali 51	DC-4000	18.1	18.0	3.5	35	78	65	4.5	1.49
Gali 5	DC-4000	20.6	18.0	3.5	35	103	65	4.4	1.49

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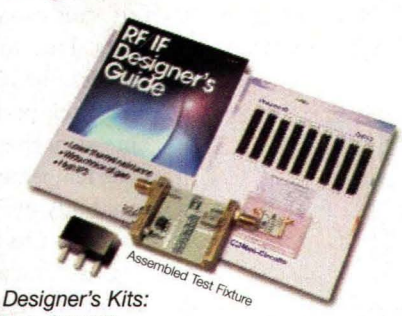

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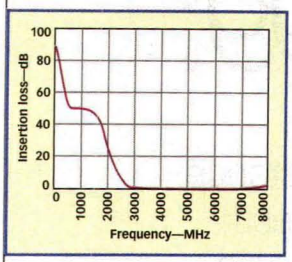
Highpass Filters Cut Off 0.6 To 3.0 GHz

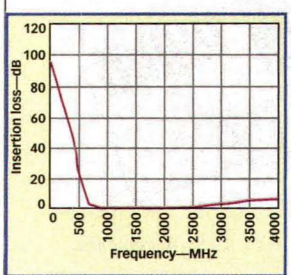
JACK BROWNE
Publisher/Editor

These ceramic highpass filters offer cost-effective, drop-in solutions for a variety of applications requiring attenuation of signals below a given cutoff frequency.

Highpass filters are simple in function—passing signals without affecting them beyond a certain (cutoff) frequency and sharply attenuating signals below that frequency—but often difficult to realize in practice. By leveraging their high-performance Blue Cell™ low-temperature-cofired-ceramic (LTCC) technology, however, the engineers at Mini-Circuits (Brooklyn, NY) have created the HFCN line of seven-section highpass filters with cutoff frequencies from 600 to 3000 MHz at a fraction of the cost of traditional highpass filters. The compact components (supplied in a standard 1206 EIA package) are well suited for subharmonic attenuation, breakdown-voltage protection, passband matching, and as DC blocks.



1. The measured insertion loss of the HFCN-650 filter demonstrates the familiar highpass response characteristics with sharp attenuation of signals below cutoff.



2. The measured insertion loss of the HFCN-2700 highpass filter shows the increasing attenuation of signals below the 2.7-GHz cutoff frequency.

Traditional highpass filters incorporate discrete components to create desired resonances. Unfortunately, such components are subject to variations in electrical characteristics and additional factors, such as inconsistent placement on a printed-circuit-board (PCB) layout, can cause further performance variations. Due to their monolithic-like LTCC process, the HFCN filters provide extremely repeatable performance and in fact offer guaranteed performance levels to 9 GHz, with good temperature stability and power-handling capability of 7 W. The LTCC filters are also insensitive to electrostatic discharge (ESD).

Examples of the HFCN highpass filter line include the models HFCN-650 **(Fig. 1)**, HFCN-740, HFCN-2100, and HFCN-2700 (**Fig. 2**), with the cutoff frequency (in MHz) of each unit denoted in the model number. At the cutoff frequency, an input signal is attenuated by 3 dB, with attenuation decreasing beyond that frequency. The HFCN-650, the lowest-frequency model in the line, for example, has a cutoff frequency of 650 MHz and is nominally designed for a passband of 850 to 2490 MHz where signal attenuation is minimal. The highpass filter provides 20-dB rejection of signals at 480 MHz increasing to 40-dB rejection at 390 MHz. Above the cutoff frequency, the passband insertion loss is low, at less than 1.3 dB from 850 to 2000 MHz and less than 2 dB from 710 to 2490 MHz. The model HFCN-740 has a cutoff frequency of 740 MHz and nominal passband of 900 to 2800 MHz. Rejection is at least 40 dB at 430 MHz and insertion loss is less than 1.3 dB from 900 to 2200 MHz.

At the high-frequency end, the model HFCN-2700 offers a cutoff frequency of 2700 MHz with nominal passband of 3150 to 7550 MHz. It exhibits better than 40 dB rejection at 1600 MHz and less than 1.3 dB insertion loss from 3150 to 6500 MHz. Insertion loss rises to just under 2 dB at frequencies to 7550 MHz. The model HFCN-2100 offers a cutoff frequency of 2100 MHz with nominal passband of 2500 to 6000 MHz. It rejects signals at 1530 MHz by 20 dB and at 1050 MHz by 40 dB. Insertion loss is less than 1.3 dB from 2500 to 5000 MHz and less than 2 dB from 2200 to 6000 MHz. All HFCN filters are designed for operating temperatures from –55 to +100°C. P&A: $0.99 (HFCN filters, 1000 qty.) and $79.95 (K1-HFCN kit of 40 filters); stock. **Mini-Circuits, P.O. Box 350166, Brooklyn, NY 11235-0003; (718) 934-4500, FAX: (718) 332-4661, e-mail: sales@minicircuits.com, Internet: www.minicircuits.com.**

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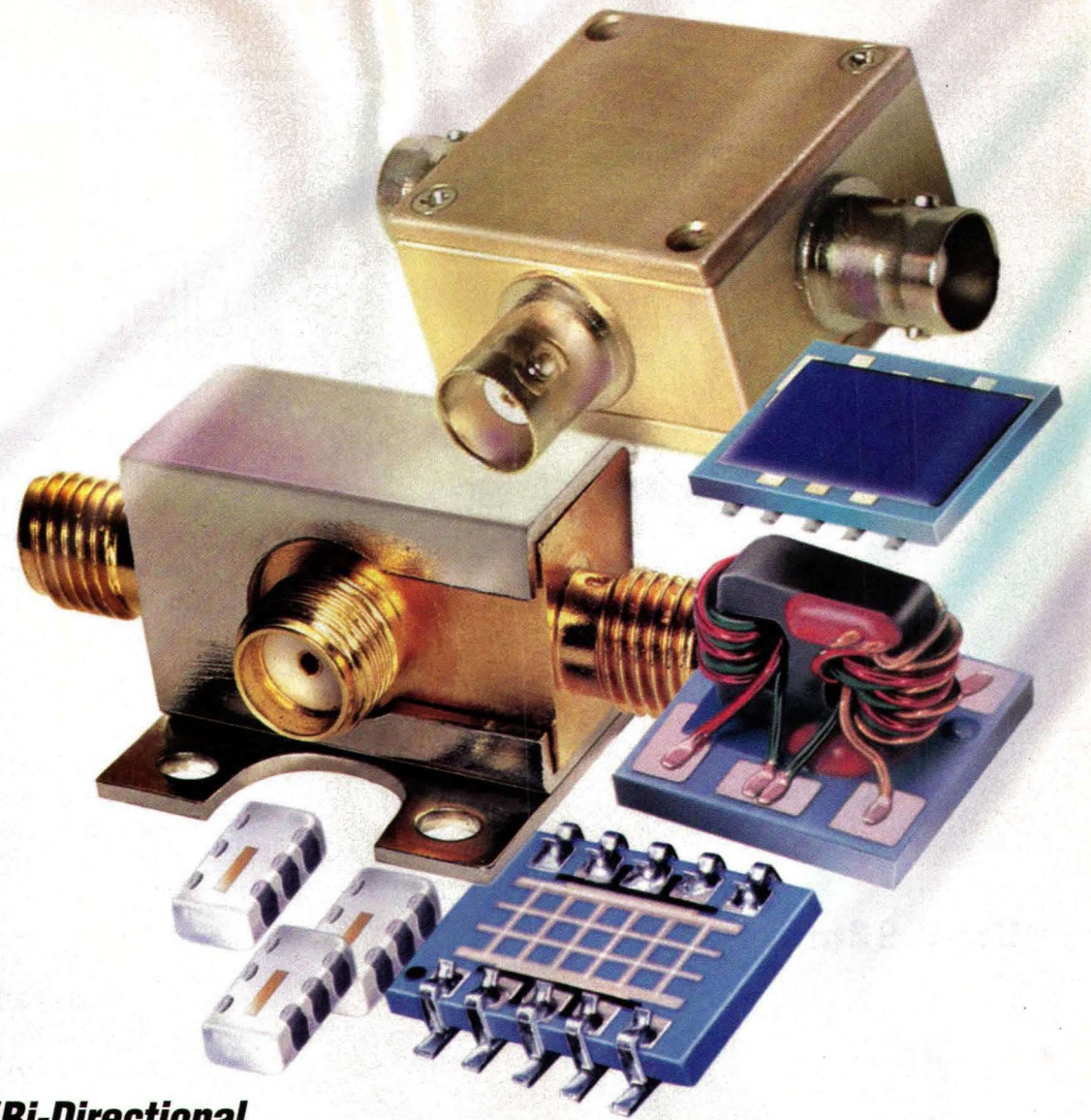

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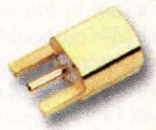
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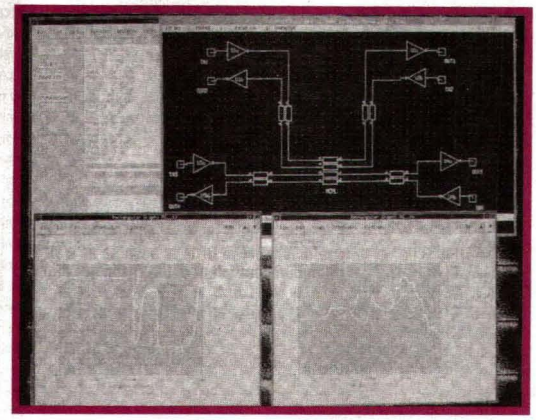
looking back



ALMOST 11 YEARS AGO, microwave design engineers had a version of the popular SPICE simulation program to call their own, as Compact Software (Paterson, NJ) introduced their Super-Spice simulation time-domain program for UNIX workstations.

next month

Microwaves & RF April Editorial Preview
Issue Theme: Wireless Technology

News

Wireless technology represents a broad area of coverage. A review of the product introductions and technical presentations made at the recent Wireless Systems Design Conference & Expo (March 8-10, 2004, San Diego, CA) offers an excellent opportunity for summarizing the current state of the art in wireless technologies, including where industry companies stand with emerging technologies such as UWB transmissions and some novel approaches to achieving bandwidth efficiency through UNB modulation methods.

Design Features

Embedded GPS radios can now be found in many high-end vehicles and cellular telephones, as well as in PDAs and other portable devices where location tracking is required. To help those interested in designing GPS radios, authors from Agilent Technologies provide practical guidelines for developing a GPS radio based on low-noise SiGe semiconductor technology.

Additional technical articles in April will include the conclusion of a four-part article series on FIR digital filters, a study of multipath fading phenomena in terrestrial communications systems by one of the FCC's leading engineers, and techniques for estimating multiple carrier interference and noise power ratio (NPR) in communications systems.

Product Technology

April's Product Technology section supports the wireless technology theme with a report on the state of the art in DDS IC technology. By including a DDS core, RF prescaler, DAC, phase detector/charge pump, and a differential clock driver on a single chip, engineers have access to clean, programmable signals to 2.7 GHz. April also offers a single-card, PXI-based vector signal generator covering 250 kHz to 2.7 GHz, a compact front-end module that combines FBAR filter technology with E-pHEMT power amplifier technology, and a versatile transceiver IC that works for all three 802.11a/b/g WLAN systems.

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2000 - 4000 MHz	1.5	4952 - 100 X
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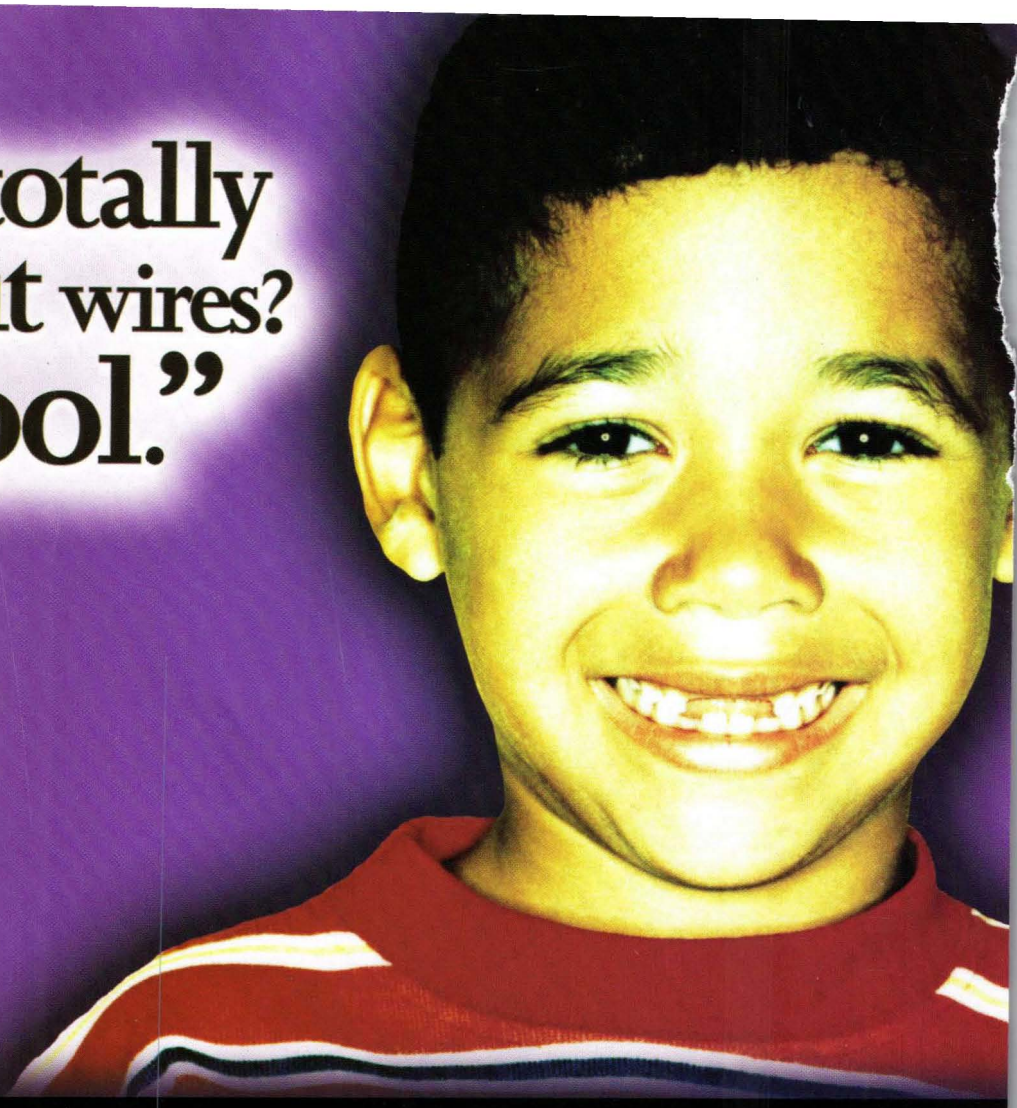
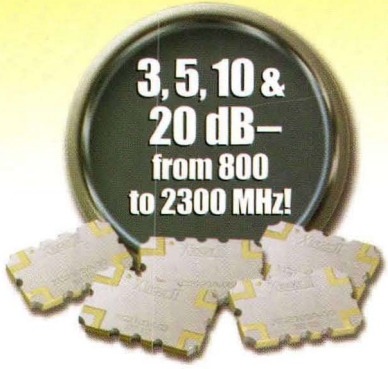


Introducing Xinger II — the next generation of components for the next generation of wireless systems.



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